6889-EE-02 ②

AD-A252 817

IEEE
Microwave Theory and Techniques Society
C.&S. Italy, France and Germany MTT Chapters

1992 International Workshop
Millimeter Waves

27-24, 1992

Capitano del Popolo,
Orvieto, Italy

Organized by:

Central and South Italy MTT Chapter
MTT Chapter
Germany MTT Chapter
University of Perugia
p.a., Orvieto

Sponsored by:

CNR, Italy
CNET, France
USARDSG European Research Office, UK
IEEE Microwave Theory and Technique Society
University of Perugia, Italy
Coaxial Italiana S.p.a., Italy
Telco S.p.a., Italy
Alenia S.p.a., Italy

92 7 10 042

92-18157

Technical Program Committee
P. Lampariello, University of Rome La Sapienza, Workshop Co-Chairman
R. Sorrentino, University of Perugia, Workshop Co-Chairman
V. Fouad-Hanna, CNET, France
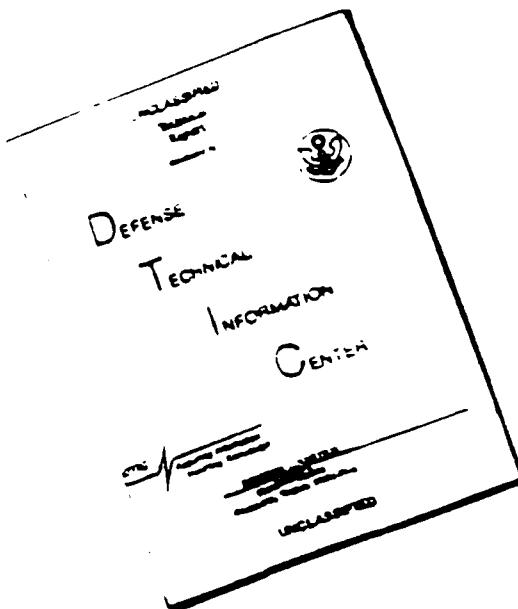H. L. Hartnagel, Technische Hochschule, Darmstadt, Germany

Organizing Committee

R. Giorgini, Itelco S.p.a.
V. Alleva, Itelco S.p.a.
S. Monelli, Itelco S.p.a.
S. Catena, University of Perugia

Secretariat
V. Papini, University of Perugia
L. Castellani, University of Perugia
C. Paolotti, University of Rome "La Sapienza"

DISCLAIMER NOTICE

THIS DOCUMENT IS BEST QUALITY AVAILABLE. THE COPY FURNISHED TO DTIC CONTAINED A SIGNIFICANT NUMBER OF PAGES WHICH DO NOT REPRODUCE LEGIBLY.

ƊƐ

1992 International Workshop on Millimeter Waves

April 22-24, 1992

Palazzo del Capitano del Popolo
Orvieto, Italy

Organized by:

Central and South Italy MTT Chapter
France MTT Chapter
Germany MTT Chapter
University of Perugia
ITELCO S.p.a., Orvieto

Sponsored by:

CNR, Italy
CNET, France
USARDSG European Research Office, UK
IEEE Microwave Theory and Technique Society
University of Perugia, Italy
Contraves Italiana S.p.a., Italy
Itelco S.p.a., Italy
Alenia S.p.a., Italy

Technical Program Committee
P. Lampariello, University of Rome La Sapienza, Workshop Co-Chairman
R. Sorrentino, University of Perugia, Workshop Co-Chairman
V. Fouad-Hanna, CNET, France
H. L. Hartnagel, Technische Hochschule, Darmstadt, Germany

Organizing Committee

R. Giorgini, Itelco S.p.a.
V. Alleva, Itelco S.p.a.
S. Monelli, Itelco S.p.a.
S. Catena, University of Perugia

Secretariat
V. Papini, University of Perugia
L. Castellani, University of Perugia
C. Paolotti, University of Rome "La Sapienza"

FOREWORD

Welcome to third edition of the International workshop on Millimeter Waves, being held at the *Palazzo del Capitano del Popolo* in Orvieto. After the encouraging success of the previous editions at the University of Rome "Tor Vergata" in 1986 and at the University of Perugia in 1989, it was decided to organize a new workshop as a cooperation between the Central and South Italy Chapter with the France and Germany Chapters of the IEEE Microwave Theory and Techniques Society.

The city of Orvieto has been chosen as the venue for this event not simply because it is located midway between Rome and Perugia, but mainly for its unique charm due to its suggestive medieval appearance and the rich artistic patrimony dating back to the Etruscan period. The Palazzo del Capitano del Popolo is no doubt an extraordinay combination between a hystorical medieval *palazzo* and a modern and efficient Congress Center.

ITELCO S.p.a. enthusiastically agreed to sponsor the workshop and has been a fundamental partner in the organization. The social program is also very attractive, including a very nice concert of two Canadian artists, T. Goudie and D. Lawton.

We would like to thank all the invited speakers who kindly agreed to participate in the workshop and the many others who have worked for its success. We are deeply grateful to Professor Carrara, one of the *fathers* of microwaves in Italy, who accepted our invitation to welcome the participants at the opening session.

Paolo Lampariello and Roberto Sorrentino

Workshop Co-Chairmen

CONTENTS

MILLIMETERWAVE APPLICATIONS AND TECHNOLOGY TRENDS

H. Meinel

Deutsche Aerospace • Germany

1992

INTERNATIONAL WORKSHOP

on

MILLIMETER WAVES

Palazzo del Capitano del Popolo

Orvieto, Italy

MILLIMETER WAVE APPLICATIONS
AND TECHNOLOGY TRENDS

Holger H. Meinel

Deutsche Aerospace AG
Leopoldstrasse 175
D-8000 München
Germany

ABSTRACT

The advantages and new application areas being offered by millimeter-wave systems have been known for a long time. But only the advent of reliable and efficient solid-state sources as well as the progress in low-cost integration procedures, i.e. Hybrid- and Monolithic-Integration Techniques, respectively, has made millimeter-wave technology mature, opening a wide field of employment areas today. It is widely accepted, that the solution to traffic congestion, being a serious problem worldwide, does not lie in the construction of more roads, but in the application of advanced technologies to control the flow of vehicles. The efficiency of the already existing infrastructure will be increased, incorporating RTI (road traffic informatics) in the millimeter-wave range.

APPLICATIONS

Three major application areas have to be distinguished in this field today, communications, radar, and automotive traffic control, the later combing the earlier two. Apart from the unique propagation behaviour, millimeter-wave applications take advantage of the highly directive nature of the millimeter-wave beam, i.e. good angular resolution can be achieved with moderately small antennas, thus millimeter-wave systems are small in size and light in weight, compared to their microwave counterparts. The lack of frequencies in the microwave bands commonly used today, is an additional important reason.

COMMUNICATIONS

Atmospheric propagation effects strongly influence considerations related to the application of millimeter-wave transmission. Normally such systems will be operated in the atmospheric windows around 35 and 94 GHz; for specific applications, the maximum absorption range around 60 GHz will be taken.

The employment of the 40 - 50 GHz band for EHF military satellite up-links has been very important for the development of millimeter-wave technology. This frequency band offers increased bandwidth over the UHF and SHF bands being in current use today and permits the construction of small terminals: thus covert operation for various military platforms is achieved. The SKYNET 4 series of satellites in the U.K. and the MILSTAR program in the US are such applications.

As new services emerge for LAN (local area network)and PCN (personal communication network) systems as well as for traffic control, the demand for a new generation of high data rate radio links has fostered the worldwide development of commercial millimeter-wave POINT-TO-POINT transmission link equipment.

VIDEO-CONFERENCE-SERVICES or SATSTREAM and MEGASTREAM are application examples for high data rate radio links, while remote telephone- or 'cable'-TV-subscriber connection is another; here with special emphasize on 'low cost and effort' infrastructure enhancement. The safety control of high speed trains with velocities up to 400 km/h, like the German ICE (Inter City Express) and the French TGV (Train à Grande Vitesse) as well as LCRN-services (local cellular radio network) are mobile millimeter-wave applications.

- POINT-TO-POINT transmission links, using portable equipment can be installed very rapidly on maximum hop lengths of up to about 10 km, the equipment, due to size and weight, can be mounted easily.

The main employment area for such millimeter-wave equipment is in connecting subscribers, for example in private data -, TV -, and telephone circuits. More than 1,000 systems have become operational since the late 1970's, mostly in Japan, but also in Europe.

Besides these point-to-point and area type connections, the millimeter-wave range has been used and is further investigated for mobile communication purposes.

- Local Cellular Radio Networks (LCRN), for wideband mobile services, are a major investigation effort, for instance in England.

The demand for mobile radio has been such, that severe spectrum congestion is occuring within the today allocated bands. Thus a region of radio spectrum has being sought, which could accomodate the future LCRN bands. The application of millimeter-waves could be the solution.

RADAR

The widest spread employments of millimeter-waves are found in radar, having fostered the development of millimeter-wave technology during the last two decades. Seekers for terminal guided submunitions or battlefield and terrain mapping radars are military application examples, while the cont :tfree measurement of distances or velocity in industry and traffic control are typical civilian employments.

The most important application in this area were seekers for terminal guided submunitions. Besides stringent performance and low cost constraints a high production volume was required; thus the world-wide seeker development has stimulated millimeter-wave development efforts and secured todays technology status. Such seekers, engineered to production status, are available from various manufacturers.

Millimeter-wave radar is advantageous for applications, where high resolution is a major requirement. Short range surveillance radars, like the Helicopter Obstacle Warning Radar in the lower frequency regime (40 - 60 GHz), is an example of this type. Wire like obstacles with diameters as small as 3 mm have been detected under flight conditions. Being a military necessity at first, helicopter obstacle warning today is an important civilian application, concerning the safety of airport commuter traffic.

TRUE MOTION SENSORS Future traffic control systems require true ground speed information for ABS, ASC or car navigation systems. Using Doppler systems at 24 and 61 GHz, respectively, it has been shown, that at least two independent sensors have to be employed, - looking forward and backward - in order to correct the systematic measurement error due to vehicle movements. Several sensors have been developed at different Institutes and Companies in cooperation with the major car manufacturers in Europe.

AUTOMOTIVE COLLISION AVOIDANCE RADAR The use of radar for automotive collision avoidance or intelligent cruise control is not new. Many companies, institutions and laboratories experimented with various millimeter-wave frequency pulse- and/or FM-CW radar systems in the past two decades. Having small size antennas to be fitted into the front of a car such radars are built for short range use only, up to 200 m maximum have to be and can be achieved easily, even under adverse weather conditions.

REFERENCES

Applications:

[1] Van der Forst, A. : mm-Wave Atmospheric Propagation and System Implications. 16th EuMC, Dublin, Sept 1986, 19 - 30.

[2] Meinel, H.: System design, applications and development trends in the millimeter-wave range. 18th EuMC 1988, Stockholm.

[3] Ron Schneiderman: Millimeter-waves find commercial markets, Microwaves & RF, July 1991, 35-39.

[4] Meinel, H.: Millimetre-wave technology advances since 1985 and future trends. IEEE Trans. MTT 39, No. 5, May 1991, 759 - 767.

Communications:

[5] Dawson, T.W.G., et al.: Skynet 4 defense communication satellite project. MM (82), London Oct. 1982, 94-103.

[6] Brigginshaw, P. M.; Restarick, A. S.: 60 GHz covert communication link in binoculars. IMM '86 Conference, Brighton Jun. 1986.

[7] Plattner, A.; Meinel, H.: A 60 GHz hand-held LPI transceiver. 12th EuMC, Helsinki, Sept 1982, 540-594..

[8] Meinel, H.: The current status of mm-wave communcation systems, melecon '89 conf., Lisbon, Portugal, March 1989, 680-686.

[9] Dooi, Y.; Kurematsu, H.; Furuhama, Y.: A 50 GHz band millimeter-wave radio system, Links for the future, Science systems and services for communication, IEEE publication: CH 2028 - 9/84, 955-958.

In industry millimeter-wave radar can be employed in any situation, where movements, length or height have to be measured in a contactfree manner. Doppler sensing is taken to gain information for automatic process control. An example of the later is the top-dead-center measurement system for combustion engines, while the controlled glue application on windows in car manufactoring is an example of the earlier.

RECONNAISANCE

Millimeter-wave technology has matured and operational systems are deployed in the military arena; thus millimeter-wave reconnaissance has become an urgent necessity today. Different systems, including millimeter-wave superhet- and DIFM - receivers, have been built.

TRAFFIC CONTROL

Millimeterwave techniques offer the means of sensing vehicles and communicating with them, while they have the advantage of being immune to changing weather conditions. Four major areas of applications have to be distinguished: Automatic Debiting Systems (ADS) and Road traffic Informatics (RTI) are communication type applications, while true motion - and collision avoidance sensors are radar type approaches. Especially millimeter-wave systems have found an increasing interest for such purposes due to their specific advantages: to be small in size and light in weight. The lack of frequencies in the microwave bands commonly used today, is another very important reason. The frequency allocations for European road transport applications, as suggested by CEPT, clearly identifies the importance of millimeter-wave techniques in this field.

ADS TELEPASS is an ADS installed along the Autostrada Milan - Naples in Italy. Communication is achieved at 5.72 GHz between a roadside transceiver at the tollplaza and a transponder within the vehicle. Such an ADS approach will also be taken to reduce city traffic significantly, e.g. in Cambridge, England, as city entrance debiting is the second major ADS application. In the future such systems will be operated in the 60 GHz range.

RTI A major concern of RTI is vehicle-to-vehicle and vehicle-to-roadside communication. An example of this type is the DACAR system within the DRIVE programme. Communication occurs at 61 GHz over a taken prototype bandwidth of 500 MHz. DPSK (Differential Phase Shift Keying) with error detecting and correcting coding was applied. This leads to a bit error rate (BER) of 5 x 10 expl- 10) after demodulation, as multiple tests trials, being carried out in May/June '91 have shown. Further systems, like the improved German AVES (AEG Verkehrs Erfassungs System) at 61 GHz are under test and development, respectively, for the same purpose.

[24] Stammler, W.; Boheim, M.; Linss, W.: mm-wave sensor for traffic data acquisition systems, Proc. Int. Microwave Symp. (Sao Paulo, Brazil), July 1989. 24-27.

[25] Maege, B.; Hofmann, K.-W.; Happe, A.: Determining TDC in fired internal combustion engines, Technisches Messen 53 (1936), No. 1 Jan. 1986, 25-27.

Reconnaisance:

[26] Newton, B.H.; Dodds; A.F. Nugent, D.M.; Body, D.H.: A sensitive K-band receiver for ESM applications, MM (86), Brighton June 1936.

[27] Meier, P.J.; Breuer, K.D.; Cohen, L.D.; Worontzoff, N.; Lepore, J.; Gunther, J.: Channelized receiver covering 26 to 60 GHz with planar integrated circuit components, Eaton Corp. AIL Div., Print 81 - 17.

Traffic control:

[23] Panel discussion on: "Microwaves and Optronics in Vehicular Technology", MIOP '90, April 24-26 1990, Stuttgart, Germany.

[29] Workshop on: "Advanced car electronics and future traffic control systems related to microwaves". 21st EuMC 1991, Sept. 13 1991, Stuttgart, Germany, Workshop Proc., 97-153.

[30] Glathe, H.-P.: Contribution to a future European traffic structure - progress report. 8th Int. Conf. on Automotive Electronics, Oct. 1991, London, England. 112 - 119.

[31] Zandbergen, A.: Evaluation of different transmission media for use in automotive debating systems, DACAR Conf., May 1991, Roma, Italy.

[32] autostrade S.p.A.: TELEPASS, toll-payment without stopping, central department for information, Florence, Italy - 1991.

[33] The Cambidgeshire County Council: Congestion Metering in Cambridge City. Information Broschure 1992.

[34] Blythe, P. T.: The use of transponder technology in road traffic control, 8th Int. Conf. on Automotive Electronics, Oct. 1991, London, England, 130 - 134.

[35] Moss, C. R.: The DACAR millimetre wave vehicle communications link, Workshop 21st EuMC, Stuttgart 1991, in (29), 112 - 121.

[36] Fischer, H.J.: Digital beacon vehicle communication at 61 GHz for interactive dynamic traffic management, 8th Int. Conf. on Automotive Electronics, Oct. 1991, London, England, 120 - 124.

[37] Kehrback, J.; Heidrich, E.; Wiesbeck, W.: Planar microwave Doppler-sensors for car speed monitoring Workshop 21st EuMC, Stuttgart 1991, in (29) 129-134.

[10] Watanabe, S.; Kasuga, O.; Kaneko, K.: General purpose 50 GHz Transmitter-Receiver. Links for the future. Science systems and services for communication. IEEE publication: CH 2023 - 9/84. ~51-954.

[11] Ogawa, K.; Ishizaki, T.; Hashimoto, K.; Satakura, M.; Uwano, T.: A 50 GHz GaAs FET MIC transmitter/receiver using hermetic miniature probe trasitions. IEEE MTT-Trans., Vol. 37, No. 9, Sept. 1989. 1434-1441.

[12] Barnes, B.C., Bates, R.N.; Clarke, I.M.: Integrated E- plane millimeter-wave 29 GHz transceiver, MSN & CT, Sept. 1989. 99-103.

[13] Barnes, B.C., O'Sullivan, P.J., Pote,A., Pilgrim, M., Wilson, P.G.: Monolithic circuits for mm-wave multichannel TV system, MIOP '90, Stuttgart, Germany. April 24-26, 1990. 426-430.

[14] Huish, P.W.; Johnson, I.T.; Li, T.A.; Nicholls, M.J.: mm- wavelength microcellular mobile radio systems - a solution to spectrum congestion for high user densities. Int. conf. on Mobile Radio, Cambridge, UK, Dec. 1935.

[15] Thomas, H.J.; Siqueira, G.L.; Cole, R.S.: Millimeter-waves shi / merit for mobile Cellular Radio, Microwaves and RF, Febr. 1937, 67-70.

[16] Bansebti, M.; Davies, R.; McGeehan, J.P.; Beach, M.A.; Rickard, D.C.: Short range propagation measurements and modelling at 60 GHz for LANs, MIOP '90, Stuttgart, Germany, April 24-26, 1990, 431-437.

Radar:

[17] Wilke, K.-H.: Microwave sensors for intelligent ammunitions MILTECH, No. 5, 1936 May 1936, 32-42.

[18] Burnett, C. E.: 94 GHz integrated monopulse radar demonstrator, IEEE MTT-S Symp., Las Vegas July 1937.

[19] Williams, M. B.; Clarke, I. M.: Radar transceiver for smart-munitions applications, MSN & CT Dec. 1937, 43 8g.

[20] Greenhalgh, D.: Millimeter wave radar for short range air defense. MM86.

[21] Bischoff, M.; Dreier, J.; Dreher, H.; Patzwahl, M.: An airborne solid state U-band pulse compression radar, MM (83), Wembley July 1933, Symp. Dig. 455 -459.

[22] Meinel, H.; Wippich, H.G., et alt.: LPI-radar for helicopter obstacle warning. AGARD Conf. Proc. No. 359, Monterey, CA, 1934, 17-1 to -7.

[23] Anthouard, P.: Millimetric wave ROMEO for obstacle avoidance (in French). Conference de Radar, Paris 1933, Communication AV-3, 192-196.

[38] Lissel, E.: Geschwindigkeits- und Wegsensor nach dem Mikrowellen-Doppler-Prinzip. VDI Berichte No. 687, 1988, 257 - 275.

[39] Lowbridge, P.L.: Brigginshaw, P.M., Kumar, B.: Millimetre wave technology for collision avoidance and cruise control. 8th Int. Conf. on Automotive Electronics. Oct. 1991, London, England, 150 -154.

[40] Stove, A.: 80 GHz automotive radar. 8th Int. Conf. on Automotive Electronics, Oct. 1991, London, England, 145 - 149.

[41] Rolland, P.A.; Haese, N.; Benlamlih, M.: Quasi optical technology for mm wave car electronics. Workshop 21st EuMC 1991, Stuttgart, in [29], 141-146.

[42] Raffaelli, L.; Stewart, E.: mmW monolithic components for automotive applications. Alpha Industries Reprint, 1992.

LPI - COMMUNICATION
Performance Data

MANUFCT.	FREQ.	OUTPUT POWER	MODULATION	DATA RATE	HOPLENGTH	YEAR
	GHz	mW		Mb/s	km	
Hughes	33	100 (max)	FM	5	8	1975
M/A-COM	33	70	NCFSK	20		
Philips	64	100	FM		4	1975
WJ	60	70	FM	BW 5 MHz		
NOSC	60	50	FM		0.1	1955
AEG	55	10	FM	BW 200 kHz	1.5	1955
TRW	70	10 (max)	PCM/FM	BW 150 kHz	7.3	1977
MARCONI	60	50	AM			1955

LPI : Low Propability of Intercept

SHORT-RANGE RADAR SYSTEMS

COMMUNICATION SYSTEMS

TRAFFIC CONTROL

MILSTAR

Deutsche Aerospace

BLOCK DIAGRAM OF A 50 GHz TRANSMITTER AND A RECEIVER

TRANSMITTER   RECEIVER
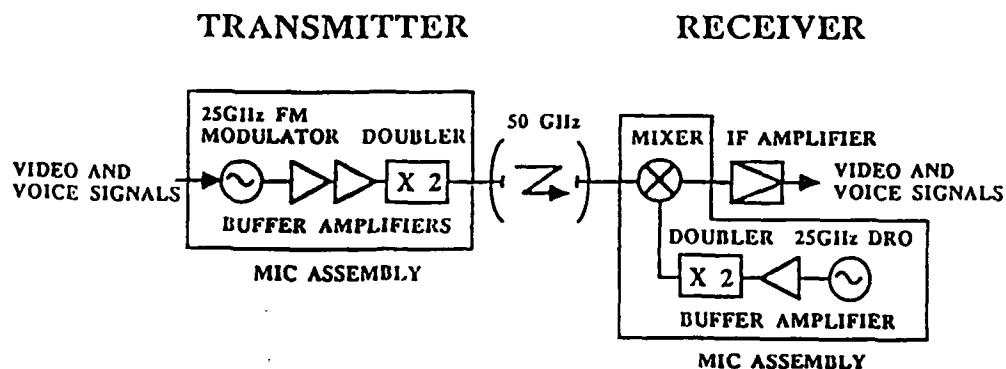
MATSUSHITA
Electric Industrial Company, Ltd.

MATSUSHITA
Electric Industrial Company, Ltd.

Major characteristics of the 50 GHz communication system

Frequency	50.46 GHz
Transmitting power	10 dBm
Noise figure	13 dB
Frequency stability	±100 ppm (-20–60 °C)
Modulation	FM
Frequency deviation	16 MHz pp
Intermediate frequency	960 MHz
IF bandwidth	27 MHz
Video signal bandwidth	4 MC
Audio signal bandwidth	15 KHz

Wicol 38

38 GHz Wireless Link
for Digital Data Transmission
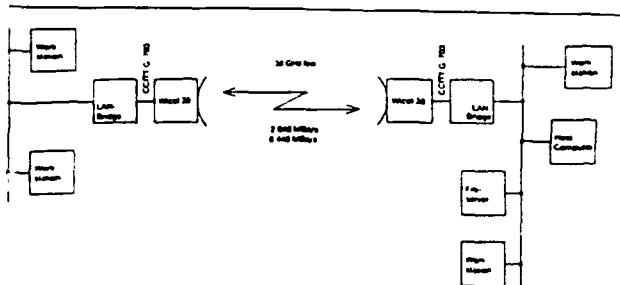
technical data:

Application	wireless interconnection of data networks e.g. Ethernet, IEEE 802.3, Token Ring FDDI. Wicol 38 is equipped with an interface corresponding to CCITT G.703 and may thus be easily used in telecommunications area
Frequency range	in the range 37.5 GHz to 39.5 GHz
Frequency tuning	channel selection by means of incorporated keyboard
Antenna beam width	approx. 2°
Bit rate	2.048 Mbit/s or 8.448 Mbit/s
Interface	corresponding to CCITT G.703
Hop length	high transmission reliability up to 4 km
Power supply	70 V to 60 V DC 230/110 V AC optional
Power consumption	approx. 50 W
Temperature range	-20 C to +50 C (other ranges optional)
Dimensions	Width approx. 290 mm Height approx. 290 mm Depth approx. 450 mm
Weight	approx. 18 kg

Master Antenna feed to properties where trees etc. block radio path

System Concept

-11 GHZ MULTICHANNEL TV SYSTEM FROM BTRL. MARTLESHAM HEATH, ENGLAND

✦ Deutsche Aerospace

Manufact.	Type	Freq. GHz	Output power mW	Modulation	Data Rate kbits/s	Hoplength km
Alc.-Th.	TM 440	40	50	FM	BW:15MHz	15
BTRL		29	120	FSK/PSK	8k/70k	10
Telefunken	Wicol 38	38			8,000	4
NEC	PL 50	50	15	FSK	1,544	2
FUJITSU	FC2160	50	10	FM/FSK	20,000	2
MATSUSHITA		50	10	FM	BW:15MHz	
SONY		50	20	FSK	30,000	

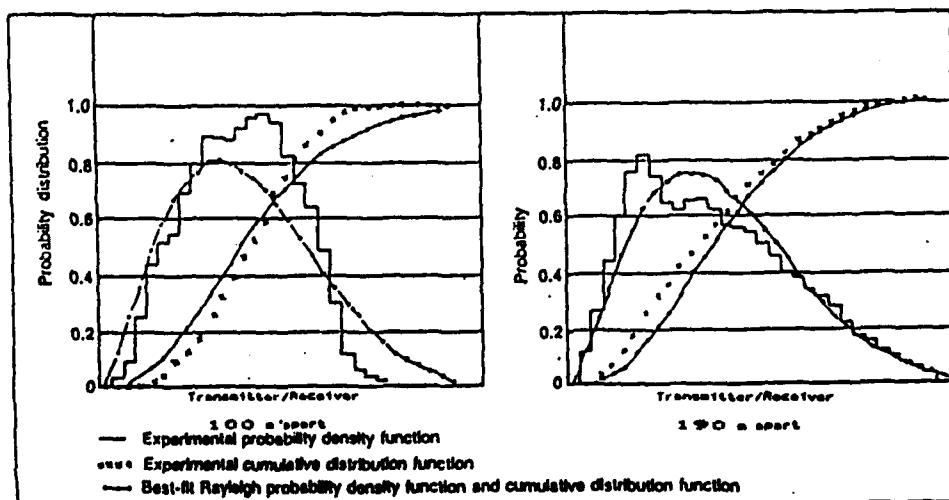**POINT-TO-POINT Millimeter-Wave Communication systems**
Performance Data
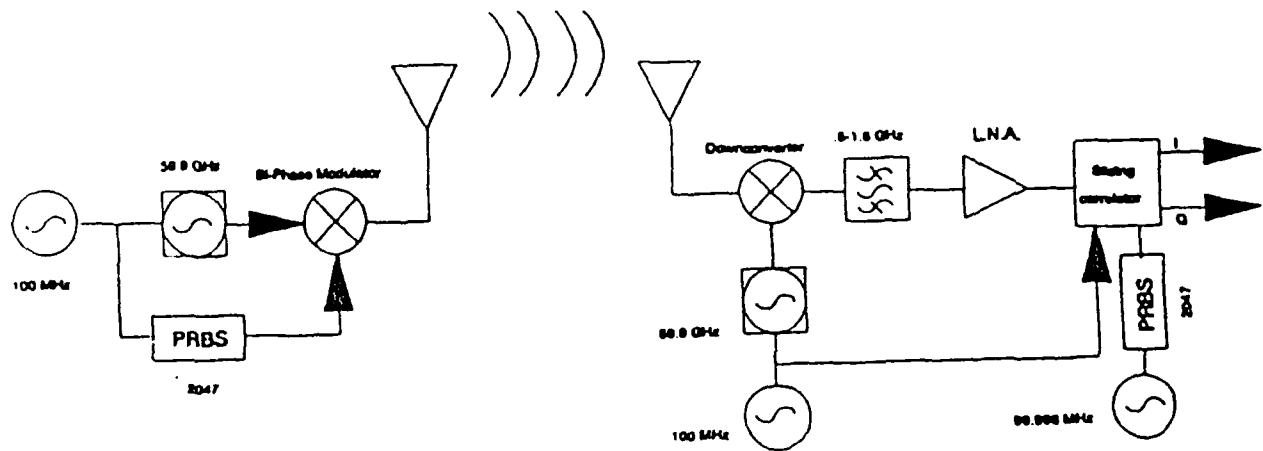
Sounder Diagram

---

60 GHZ LAN PROPAGATION MEASUREMENTS
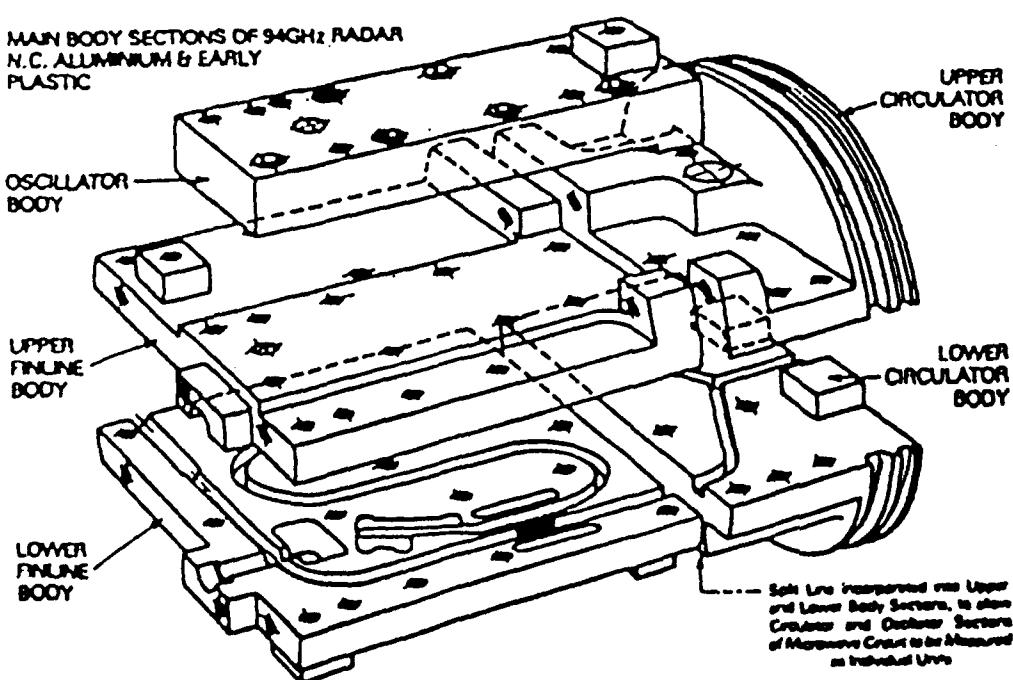FROM: UNIVERSITY OF BRISTOL, UK

Radio and Radar Systems Division

AEG

---

100 a apart

— Experimental probability density function

···· Experimental cumulative distribution function

— Best-fit Rayleigh probability density function and cumulative distribution function

from: H.J. Thomas, S.L. Siqueira, R.S. Cole
UNIVERSITY COLLEGE LONDON,
London, England

---

60 GHz PROPAGATION IN AN URBAN ENVIRONMENT
Probability density function

Manufact.	Freq.	Output	Modulation	Bandwidth	Runlength
	GHz	Power mW		MHz	m
Univ. Col. Ldn.	55	60		5	150
Univ. Bristol	60	20	PRBS	2	L-O-S 200

Experimental Millimeter-Wave System for Mobile Braodband Services (MBS) Performance Data

PHILIPS MICROWAVE

MAIN BODY SECTIONS OF 94GHz RADAR
N.C. ALUMINIUM & EARLY PLASTIC

UPPER CIRCULATOR BODY

OSCILLATOR BODY

UPPER FINLINE BODY

LOWER CIRCULATOR BODY

LOWER FINLINE BODY

TDC measurement — Schematic diagram

Injection

Coupler

Quartz

Pre-Chamber

Photodiode

Generator

Absorber

Firing
Triggersignal

Misc Mixer

Triggergenerator
150 KHz

Signalamplitude

Processor

| 4500 | 10,2° | -0,47° |

Angle Markings
relative to TDC

| RPM | Firing
Trigger Ang | Real TDC
(dyn. meas.) |

✦ Deutsche Aerospace

| | REQUIREMENTS FOR AN OBSTACLE WARNING SYSTEM | **AEG** |
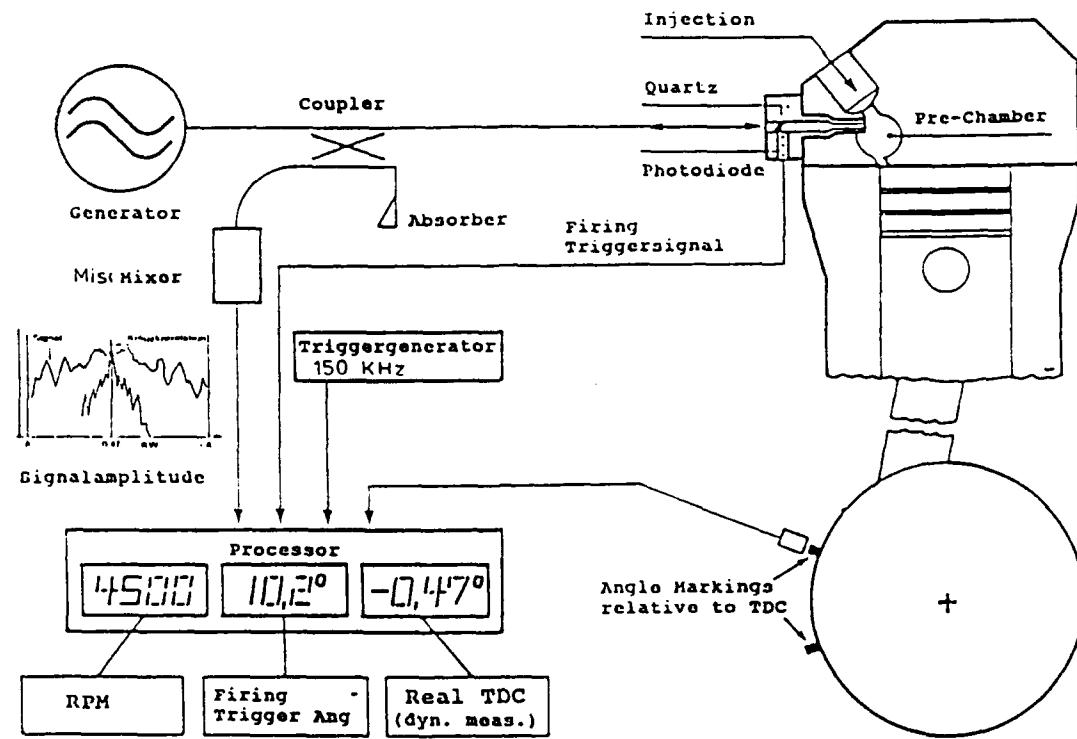
Output power	35 mW
Frequency	94 GHz
Sweep range	800 MHz
Linearity	less than 1 %
Mixer NF	8 db
Weight	less than 50 g
Price *
(housing)	less than 5 £

- DETECTION OF CABLES AND WIRES WITH DIAMETERS >3mm

- RANGE OF CA. 500m

- ALL WEATHER CAPABILITY

- LIGHT WEIGHT AND SMALL VOLUME

- COOPERATION WITH EXISTING NAVIGATION AIDS

*) polymer injection moulding process
more than 10,000 units
Cu-plating applied

Philips Microwave
94 GHz seeker data
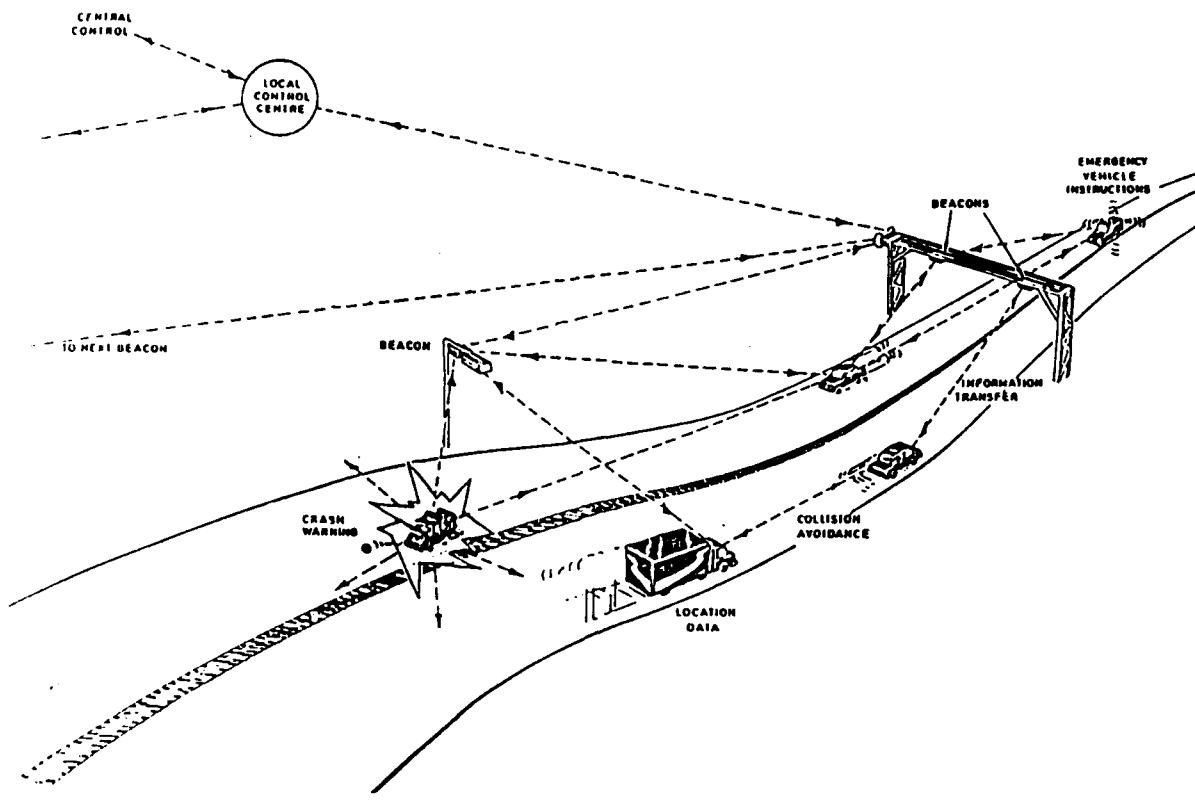
Figure 2.1:   Operational system for RTI applications

18

Deutsche Aerospace

Bands GHz	5.795 - 5.805 5.805 - 5.815	63 - 64	76 - 77
Applications	Automatic toll debiting systems	Transmissions Road ↔ V V ↔ V	Anticollision radars
Bandwidth	2*10MHz or 4 channels of 5 MHz	1 GHz Channels of 5 to 20 MHz	FM/CW : 100 MHz Pulsed : 500 MHz
Allowed power	3 dBw	3 - 16 dBw	16 - 20 dBw
Recommanded antenna gain	10 - 15 dB	10 - 30 dB	30 - 35 dB
Rate	1 - 3 Mbits/sec	Few Mbits/sec	
Modulation	FSK, PSK, ASK	FSK, PSK	FM/CW Pulsed

from: HESS, President of the CEPT frequency group - DRIVE Conference, 4 - 6/2/91

CEPT Frequency Allocations for
European Road Transport Applications
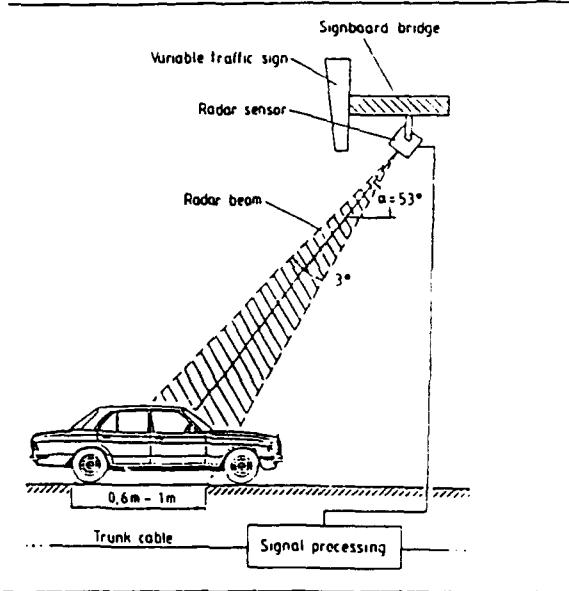
Application of AVES - Millimeterwave Radar Sensor

Deutsche Aerospace
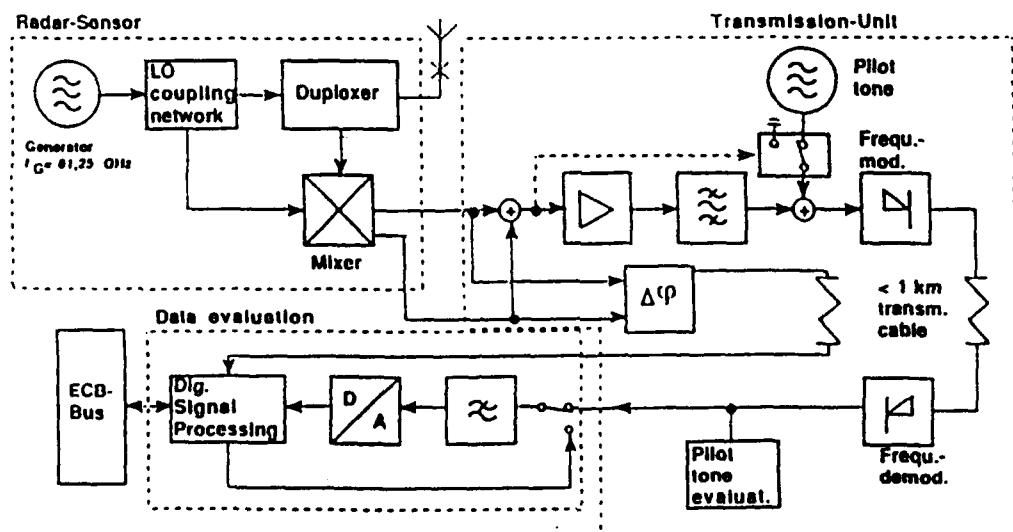
AVES — AEG VERKEHRS ERFASSUNGS SYSTEM

SCHEMATIC BLOCK DIAGRAM

|	
Output power	5 mW
Frequency (2nd harm. Gunn)	61.25 GHz
Harmonic suppression (H-irises)	> 25 dB
Antenna beamwidth	
azimuth	13 degrees
elevation	3 degrees
(asymetric lens)	
gain	25 dB
Mixer NF (Ku-Band diodes)	dB

Telefunken Systemtechnik
AVES sensor data

1) RANGE ...
CONFIRMED.

2) OPERATES SATISFACTORILY IN PRESENCE OF MULTIPATH EFFECTS.

3) OPERATES IN PRESENCE OF OTHER VEHICLES.

4) OPERATES WHEN DIRECT PATH IS OBSCURED BY OTHER VEHICLES.

5) ERROR RATE BEFORE ERROR CORRECTION $\simeq 10^{-4}$.

6) ERROR RATE AFTER ERROR CORRECTION SO LOW THAT IT IS DIFFICULT TO MEASURE (PROBABLY LESS THAN 10^{-9}).

DRIVE - DACAR - Program
Road test results

AEG

Radio and Radar Systems Group
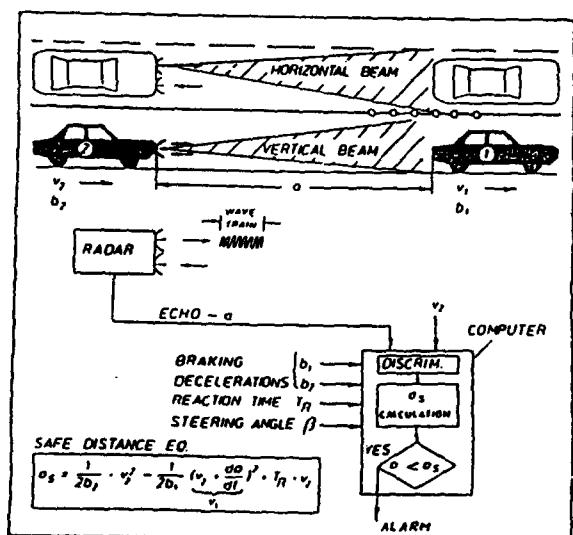
5 GHZ collision avoidance radar
rinciple of operation

TOYOTA MOTOR CORPORATION, JAPAN

PERFORMANCE DATA

STAND ALONE 'SKIN TRACKING' UNIT

FREQUENCY	49.5 GHZ
MODULATION	
FM-TRIANGULAR	75 MHZ
RATE	750 HZ
OUTPUT POWER	30 mW
MEASUREMENT RANGE	3 TO 60m
150mm/h RAIN	TO 50m
ACCURACY	5cm/h

ANTENNA DESIGN

SEPARATE TRANSMIT AND RECEIVE ANTENNA
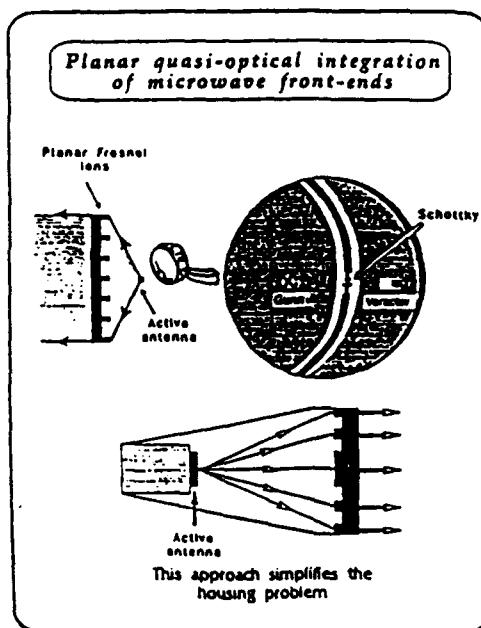2 DEGREES BEAM WITH
CROSS POLARIZED
POLARIZED 45 DEGREES TO ROADWAY

49.5 GHZ AUTOMOTIVE RADAR

Deutsche Aerospace

Planar quasi-optical integration of microwave front-ends

This approach simplifies the housing problem

Quelle: P.A. Rolland, C.H.S · Université de Lille

Abstandswarnradar
Quasi-optischer Aufbau bei 60 und 86 GHz

Deutsche Aerospace

Millimeter-wave technology is mature today

- covert communication systems
- LAN radios (PCN, MBS)

- Terminal guidance systems
- collision avoidance sensors

have been built and tested successfully

Low cost techniques

for high rate production,
hybrid and *monolithic*, are available and
prototype production has been launched

Commercial applications

have gained momentum

- RTI-Road Traffic Informatics
- LAN radio

are today's driving forces

Conclusion

---

21 EUR MICROWAVE CONFERENCE — microwave — MIOP '91

DESIGNED FOR THE INTERNATIONAL MICROWAVE COMMUNITY

THE INTERNATIONAL CONGRESS CENTRE · STUTTGART · GERMANY
9-12 SEPTEMBER 1991

WORKSHOP ON: ADVANCED CAR ELECTRONICS AND FUTURE TRAFFIC CONTROL SYSTEMS RELATED TO MICROWAVES

Year	Collision Avoidance		Doppler Sensing		Beacon Systems	
1970	Bendix	16 GHz duplex CW				
1972	RCA	10 GHz FM-CW	TI	55 GHz Speedometer		
1974	AEG-TFK	35 GHz pulsed				
	Lucas Ltd.	32 GHz FM-CW				
1975	SEL	16.5 GHz FM-CW				
1976	AEG-TFK	50 GHz pulsed	Philips	10 GHz motion sensor		
1977	Bendix	36 GHz duplex CW	AEG-TFK	35 GHz vsb-Radar		
1978	SEL	35 GHz FM-CW	Japan	Skid control sensor		
	Nissan	24 GHz pulsed/FM				
1980	Toyota/ Fujitsu	50 GHz FM-CW			Vehicular Communication	60 GHz
1982	Nissan	60 GHz pulsed/FM				
1984						
1986						
1988	Philips	94 GHz L/C Hybrid	AEG	60 GHz AVES		
1990	Uni Lille	94 GHz Quasi-optical			MCCS / SEL	60 GHz DACAR
					TST	60 GHz AVES-COM

Millimetre-wave commercial applications and technologies

As the microwave spectrum becomes more crowded the need for application of the millimetre spectrum becomes more acute. Holger Meinel from Deutsche Aerospace, Munich has collected together some of the most likely applications of this part of the spectrum and the technologies that are an essential part of the successful application.
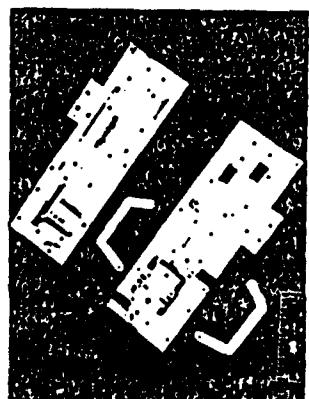

Figure 1 A 37GHz multichannel radio integrated transmitter module fabricated using M/A-COM's standard manufacturing package technique (Courtesy of M/A-COM)

Figure 2 FACOM 2160 Cross coupling filter (CCF) design (From Door et al [7]) Horizontal or vertical polarization can be selected simply by rotating the antenna by 90°

M illimetre-wave technology is available and mature today, and it is exploited in various specialist applications. In the commercial arena two major types of applications have to be distinguished: communications and traffic-control systems. The former being either point-to-point or LAN (local area network) systems, the latter incorporating radar and beacon (short hop link) systems [1].

Apart from the atmospheric unique propagation behavior [2], all mm-wave applications take advantage of the highly directive nature of the mm-wave beam, good angular resolution can be achieved with moderately small antennas, thus mm-wave systems are small in size and light in weight compared to their microwave counterparts. The lack of frequencies in the microwave bands commonly used today, especially for communication purposes, is an important reason to take advantage of mm-wave techniques.

Atmospheric propagation effects strongly influence considerations related to the application of mm-wave transmission links. Normally such systems will be operated in the atmospheric windows around 35 and 94GHz respectively. For specific applications and design reasons the maximum absorption range around 60GHz might be taken to ease the reuse of frequencies, for example.

Being the first to take advantage of the mm-wave approach the technology of early communication systems was mostly waveguide related or derived.

Waveguide integration

The VIP (Waveguide Integrated Package) technique for integrating several different waveguide components into one compact, uniplanar integrated assembly has been developed and used at M/A-Com of Burlington, MA, USA, for many years in the microwave range. It has been used as a cost-effective method for high-volume production, providing size and weight reduction without sacrificing performance. VIP technique use waveguide as the primary transmission medium and computerized NC milling combined with laser welding for fabrication.

The approach was taken to develop a low-cost compact integrated mm-wave front-end for LPI (low probability of intercept) communication for the US-Army [3]. This Ka-band

radio module was designed to be used with existing host radios.

System parameters for the host radio include 60mW output power at 70MHz an antenna gain of 41dB and 8.5dB receiver noise figure.

The integrated 37GHz transmitter assembly of this mm-wave radio module figure 1 with the Gunn/VCO-isolator source being incorporated as a 'bolt on' unit, illustrates the system complexity M/A-COM projected that the cost of VIP equipment in production at 100units/month would reduce by approximately 30%.

Planar transmission media

Hybrid and monolithic mm-wave integrated circuits have been developed, with cost, size and light weight being the major design requirements. Microstrip [4], coplanar line, suspended stripline, finline [5], and dielectric image line are the most widely used planar or 'quasi'-planar transmission media. Ease of fabrication, cost, capability to integrate a wide range of circuit functions, and specific performance features are of prime importance. Thus a tradeoff between these options based on a performance comparison is necessary in the final choice.

Beyond that, existing in-house component technology is heavily influencing this decision and thus the system design. The physical set-up for a particular application might differ significantly from company to company as seen in the following.

Based on promising measurement results customized digital distribution radios have been developed worldwide. About 700 mm-wave point-to-point radio links are already installed in Japan alone [6].

Being either usable for digital data transmission up to 20Mbit/s or for wide-band analogue signals such as colour TV, the FACOM 2160 50GHz system built by Fujitsu of Kawasaki, Japan [7], features 2 to 5km of hop length, depending on circuit performance being required.

Typical configuration is with a parabolic antenna with 40dB gain. 10mW output power and a receiver noise figure of 17dB. In order to prevent interference problems, if two or more transmitters are used in the same frequency are used in the same neighbourhood, the FACOM 2160 features a unique, but simple solution, the Cross Coupling Filter (CCF) shown in figure 2 Horizontal or vertical polarization can be selected, simply by rotating the antenna by 90°.

Conventional MIC technology on 0.2mm alumina (ferrite) substrate is used for the integrated 50GHz components to allow mass production of the RF-circuits The single-balanced mixer was implemented, using a 3dB branch hybrid and Si beam-lead Schottky-barrier diodes.

A good example for a transmission link making use of modern semicon-

ductor technology is the 50GHz system built by Matsushita Electric Ltd of Osaka, Japan [8]. In this example the output power is 10mW and the receiver noise figure is 13dB.

This system employs 3 Terminal devices only and takes advantage of hybrid MIC integration at 25GHz with doubler circuits following. The approach will reduce production costs significantly. Figure 3 depicts the assembled 25GHz FM modulator, including a dielectric resonator.

An example of a European point-to-point transmission link is the 29GHz BTL SYSTEM. Early work at British Telecom Laboratories (BTL) in Martelsham, UK, concentrated on developing a 70Mbit/s system for transmission of data and fullband video traffic. Output power here is 150mW, receiver noise figure is 8dB and antenna gain is 41dB.

Philips Research at Redhill, England, who developed the RF components of this 29GHz system, has chosen finline technology for the realization. It has emerged as a front runner for frequencies of about 25-100GHz [5]. Finline techniques achieve relatively low loss and the conductor patterns are on circuit boards, which can be made reproducible using photolithographic techniques.

The RF front end contains the integrated finline modules, to which the Gunn diode local oscillator and the transmitter sources are mounted The transmitter unit, includes a bi-phase modulator, a voltage controlled attenuator, and a monitor coupler for checking power and frequency. The receiver unit shown in figure 4, contains a low noise ba-

lanced mixer with integrated IF preamplifier, an attenuator to adjust the LO level to the mixer, and a monitor coupler for oscillator power and frequency checking. The entire RF assembly is desiccated 70 transceiver units of this type have been built.

Hybrid mm wave technology is mature today. However, MMIC technology will be incorporated into such systems as soon as it becomes available. After several years of operation, the frequency of the BTL system was reallocated to 41GHz and the entire front end will now be redesigned incorporating MMICs [10]

Local Cellular

Demand for communication services that can be provided without the need for cable connections is growing rapidly as potential users become aware of services such as cordless telephones and mobile cellular radio.

The 45GHz and 90GHz bands have been extensively used but two groups of researchers in England, namely H J Thomas et al [11] and J P McGeehan et al [12] from the Universities of London and Bristol, respectively, have investigated the 55/60GHz range for LCRN applications. Figure 5 shows the power provided by 60GHz transmission measurements carried out in Bristol.

Based on these initial measurements it appears possible to establish LCRNs (local cellular radio networks) in the mm-wave range. Thus the commercial exploitation of this greatly under utilised region of the spectrum is anticipated

So far both groups have used

Figure 3 50GHz MATSUSHITA system Assembled 25GHz FM modulator (From Ogawa et al [8])

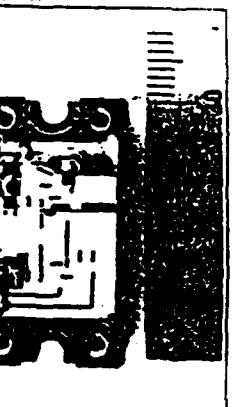Figure 4 29GHz BTL communication system integrated receiver module (Courtesy of Philips Research)

Traffic control

commercially available 60GHz equipment (standard waveguide type), see figure 6.

Traffic control

In Europe, road transportation and traffic congestion are major problems. Ever increasing demand has adversely affected traffic safety, efficiency and economy, for the future the protection of the environment is a special challenge. Pan European projects like PROMETHEUS or DRIVE are supporting solutions to control and assist the traffic flow, using Road Traffic Informatics (RTI), such as route guidance, road pricing, or incident detection, for example. Thus mm-wave systems have found an increasing interest [13]. Doppler radar systems for true motion control and traffic monitoring as well as FM-CW or pulse-radar for collision avoidance purposes, and beacon systems for road pricing and advanced road infrastructures are examples. Table 1 shows the frequency allocations for European road transport applications [14], as suggested by the CEPT. This clearly indicates the importance of mm-wave techniques in this field.

Not taking into account the very early research in this field, hybrid and monolithic mm-wave integrated circuits have been incorporated mostly. Microstrip and finline are again the two most important transmission media being employed in traffic control systems. However, quasi-optical technology, i.e. the optical integration of planar MMICs, is being studied as an option.

Future traffic control systems require true ground speed information for ABS, ASC or car navigation systems, using microwave Doppler systems for this purpose.

Fig. 5 60GHz transmission experiment: Power profile (From McGeehan [17])

Fig. 6

Figure 6 60GHz transmission experiments Schematic diagram of the sounder equipment (From McGeehan et al [13])
Figure 7 24GHz true motion sensor layout of a dual channel unit (From Kennbeck [15])

has been shown, that at least two independent sensors have to be employed, - looking forward and backward - in order to correct the systematic measurement error due to vehicle movements. Figure 7 depicts the layout of a dual channel Doppler sensor at 24GHz consisting of two T/R modules built by the University of Karlsruhe, Germany [15]. A DRO provides an output power of 6mW and the antenna has 21dB gain.

A 4 x 28 element combline array antenna is placed on the back of the substrate. For mass production, the 24GHz GaAs FET DRO uses the doubling concept with a fundamental frequency of 12GHz. The

Table 1 CEPT frequency allocation for European road transport applications

---

Figure 8 61.5GHz AVES System Schematic set-up of the mm-wave frontend (From Sammler et al [16])

self-biased single-ended mixer employs a Si beam-lead Schottky-barrier diode on a soft substrate. The complete dimensions of the microwave-frontend are 120×60× 15mm³.

The sensor module fulfills all requirements for car integration. Due to ongoing developments in high frequency satellite broadcasting systems (12GHz) the cost of the hybrid version can be expected to reduce correspondingly. Furthermore, the RF-frontend could be MMIC.

61.5GHz AVES system

61.5GHz vehicle sensors based on the Doppler-principle have been applied successfully for traffic jam avoidance or warning. The AVES system (AEG Verkehrs Erfassungs System) obtains specific traffic information from appropriate Doppler signal processing. Built by Telefunken System-Technik of Ulm, Germany [16], the chosen frequency of operation, 61.5GHz offers several advantages:
- the band is available for this application, ISM - band (Industrial, Scientific and Medical) in Europe
- due to the high frequency small antenna dimensions allow good focusing, individual lanes are illuminated only
- short range application, thus not hampered by high atmospheric absorption
- economic reuse of frequencies

Figure 9 61.5GHz AVES System Realized millimeter wave components (Courtesy of Telefunken SystemTechnik)

vehicle speed, vehicle length, and the distance between vehicles is obtained by the AVES sensor, which has been tested successfully on German Highways, outperforming a competing infrared system. The system has been approved by the German Bundesanstalt für Strassenwesen and a production contract for the first 1,000 units has been placed.

A future application of a further developed AVES sensor can be seen in vehicle-roadside communication. Together with a smart card reader such a 60GHz transceiver (hybrid or monolithic integrated) offers the possibility to establish a link between individual vehicles and a roadside traffic management network [13]. Concerning the traffic status ahead, drivers can be provided with real-time information.

Automotive radar

The use of radar for automotive collision avoidance or intelligent cruise control is not new. Many companies, institutions and laboratories experimented with various mm-wave pulse and/or FM-CW radar systems in the past two decades summarized in table 2.

Having small antennas fitted into the front of a car, such radars are built for short range use only, up to 200m, achieved even under adverse weather conditions. Present technology demands that

Figure 10 94GHz radar seeker demonstrator Cross-volume curve employing polymer-injection technique (From Binns [26])

The sensor is relatively small, approximately 150×150×150mm³, and can be mounted on existing signposts and bridges. Thus, it is not necessary to install the system underground, as is done with inductive loops, making such a sensor especially suited for mobile installations, e.g. on road-construction sites. Output power is 5mW and antenna beamwidth is 3×13°.

Hybrid integration was implemented combining finline and waveguide to reduce cost, as is the employment of a 2nd harmonic type Gunn oscillator and the implementation of Ku-band devices in the single balanced mixer. The schematic set-up of the mm-wave frontend is presented in figure 8, while the actual components are in figure 9.

Millimetre-wave applications

the antenna to be manufactured either as a chip carriers with waveguide feed. Similarly, the most efficient mm-wave oscillators utilize cavities. All the remaining front-end components can be conveniently manufactured in a single hybrid assembly using finline technology. The overall performance of such a unit is determined by the conducting housing, forming the surrounding waveguide. Plastics have generated much interest, as a low cost solution for mass production. Polymer injection moulding is one possibility, and exhibits a considerable background of process information [13]. The mass production of Compact Discs with micron surface features (a more stringent requirement compared to mm-wave radar) as achieved using a metal plated polymer injection moulding process pioneered by Philips. The performance of a 94GHz radar-seeker demonstrator realized by Philips Microwave of England are an output power of 35mW, a sweep range of 600MHz and better than 1% linearity. The mass is less than 50g.

Component values and layout design of this first demonstrator unit are not optimum for radar use, as it was necessary to have electrical access to each component for experimental purposes. However, the feasibility of this design approach was proven. Milled block technology with finline and microstrip circuits inserted was used for the realization of the RF-frontend. The housing was manufactured using metalized polymer injection moulding. The projected cost is less than £10 UK with a production rate of 45s per housing, figure 12, providing a mm-wave radar product ideally suited to the needs of the automotive industry.

The operating frequency was chosen due to military requirements, but has been changed to 76GHz. The engineer may be forced to simplify components and to combine

Quasi-optical

As mm wave systems progress to higher frequencies and complexities, the engineer may be forced to simplify components and to combine

provide a solution to this problem by combining for example, the function of an antenna and a component such as an oscillator or mixer into a single unit [19].

Examples of this approach are two anti-collision radar prototypes, for use at 60GHz and 86GHz, developed by the Universite de Lille France [14], see figure 11. The active antenna consists of a circular patch fed by a two terminal device radiating the TM₁₁ mode. Using the TM₁₁ mode current lines, a varactor for frequency tuning and a mixer diode for receiving can be incorporated as well.

Figure 11 Quasi-optical integrated radar Working Principle (variation from Rondeau [19])

functions wherever possible. Rather than scaling low frequency techniques to shorter wavelengths, a completely new approach is often needed. Quasi-optical components of a receiver frontend on a chip.

A single balanced mixer chip was developed for 60GHz and is shown in figure 14, [21] A conversion loss of 6dB and a minimum noise figure of 3.3dB was achieved. An IF amplifier using the same GaAs technology, has a gain of 29.6dB and minimum noise figure of 1.7dB at 4GHz. A 9.3Hz single balanced mixer was also fabricated with DSB conversion loss <6dB

This active antenna is placed in front of a fresnel lens.

The 60GHz pulsed Gunn radar gave excellent results for ranges up to 150m, while the 86GHz FM-CW

SIMMWIC

The application of molecular beam epitaxy (MBE) and X-ray lithography for the fabrication of monolithic integrated mm-wave devices on high resistivity silicon is very promising and has been proven [22] Discrete IMPATT and Schottky diodes fabricated from this material have shown excellent performance around 100GHz.

Hybrid integrated transmitters using the SIMMWIC (Silicon Monolithic mm-Wave Integrated Circuits) approach have delivered a maximum CW output power of 200mW at 73GHz. A completely monolithically integrated transmitting element using a planar IMPATT diode in a coplanar

IMPATT unit exhibited lower sensitivity.

Due to the inherent simplicity of the mechanical structure quasi-optical technology can be employed usefully for various industrial and commercial purposes, especially where compatible with monolithic integration.

GaAs MMIC Technology

The low-cost, very high volume production of mm-wave systems such as advanced road infrastructure systems or collision avoidance sensors will depend on the availability of mm-wave MMICs. Great efforts have been made to push the frequency limits of the key components to higher frequencies [20].

On the basis of earlier results, a new GaAs technology has been developed which allows the integration of Schottky diodes, varactor diodes and MESFETs (f_{max}=70GHz) on the same chip With this technology it is possible to integrate diode mixers, IF amplifiers and local oscillators, the key components of a receiver frontend on a chip.

Conclusion

This overview shows that mm wave technology to meet system requirements is available today. The real advent of monolithic integration techniques for production purposes has already changed the commercial area and major MMIC suppliers from the US for example, have already found partners in the European automotive industry, so the scientific discussion of such topics has become very sensitive.

However, this means that mm wave applications and technologies being initiated and pushed by military needs first are no longer "just around the corner" in the commercial arena, they have arrived.

Figure 13 SIMMWIC monolithic integrated transmitter in coplanar technology (realized at Universitet Karlsruhe, Germany)

structure was realized recently, demonstrating the feasibility. Without additional heatsinking, the output of this first unit was only 0dBm, so more work will be needed.

The integration of a coplanar Schottky-barrier diode with a micro-strip antenna yields a simple mono-lithic detector receiver chip. Developed by the Daimler-Benz Research Institute in Ulm Germany, it is shown in figure 13. The Schottky-barrier diode is terminated by a quarter-wave microstripline two dc networks for biasing and for decoupling of the received signal being connected. The Schottky diode was biased with 270mV. A receiver sensitivity of 65μV/μW was obtained at 94GHz.

References

[1] Merel H. mm wave system design and application trends in Europe Inv. Paper Ala Frascati Sept/Dec 1989 451-456
[2] Vandervorst A mm-Wave Atmospheric Propagation and System Implications Inv paper 16th EuMC Dublin Sept 1986 19-30
[3] Balzer A Loso F G Integrated frontends designed for multichannel EHF radios ESA ECP April 1987 118-127
[4] Grey T H Burret C ... mm wave (30-110GHz) mm wave microstrip technology Part1 Microwave Journal March 1986 36-44 Part 2 April May 1986 177-185
[5] Merel P J Integrated line Frascati December Part1 Microwave Journal No 85 3/93 Part 2 (bd Dec 85 (No of or 85)
[6] Fujishiro Y Abe J Microwave propagation studies EuMC 86 Dublin 113-124
[7] Goy V Kaneko H Furukawa A SDG integrated mm-wave receiver for satellite Broadcast systems and services communication IEEE Publication CH 2029 333 955-958
[8] Ogawa, Itsuki T Masumoto K Shimizu N Usami T A 50GHz compact Si-IC MIC (MMIC) for vehicular mm wave applications IEEE MTT-S Conf 86 New York
[9] Symp Dig 1992-1026
[10] Styles E C Bates R H Clarke I M Integrated mm-wave 220GHz sources 4th NSM CT Sept 1988 99-103
[11] Siemens B C et al Monolithic circuits for mm wave Commercial TV system MIOP 90 Sindelfingen FRG 1990 24-26 April 426-430
[12] Thomas H J Squira G L Goss A S mm wave front end for mobile Cellular Radio Microwaves and RF Feb 1987 67-70
[13] Brewer R McGregor J P, McNeilan Radar D C Short range processing measurements and modelling at 655 for OHz Proc MIOP 92 Stuttgart Germany 1992 24-26 April 431-437
[14] Vandervorst Advanced Car Electronics and EuMC Quasi Optical mm Wave Research Programs 23rd EuMC 1993 Sindelfingen
[15] ... 1-11-23533
[16] Rondeau FA Masse N Bertan H M ...Quasi optical technology for mm wave car electronics Workshop 21st EuMC 1991 Sindelfingen 145-146
[17] Ikeda Saito J Heinrich E Wesson K Planar microwave Doppler sensors for car speed measuring Workshop 21st EuMC Sindelfingen 1991 April 123-134
[18] Loss W mm-wave sensor for precise data acquisition systems Proc Int Microwave Symp Gas Pa - 6 Baltimore 1992 24-27 July components st[1][EuC55]-515
[19] Rondeau J D Bosse P Kazar ... Quasi-optical integrated microcircuits and mm-wave circuits and automotive mm wave application European Microwave Conference Oct 91
[20] Hirsch M J Scharf F S ... Monolithic mm-wave Integrated Circuits 1987 EuMC 139' Sindelfingen Germany 93-98
[21] Kasturi F S ... mm-wave integrated circuits Proc IEEE GaAs Symp 1989
[22] Luy J-F ... Monolithic Silicon mm wave integrated circuits Proc IEEE GaAs Symp 1991

	Collision avoidance	Doppler sensing	Beacon systems	
1970	Bendix	16 GHz doppler CW		
1972	RCA	10 GHz FM-CW	TI 35 GHz spectrometer	
1974	AEG-TFK	35 GHz pulsed		
	Lucas Ltd	32 GHz FM-CW		
1975	SEL	16.5 GHz FM-CW		
1976	AEG-TFK	50 GHz pulsed	Philips 10 GHz motorway	
1977	Bendix	36 GHz dopler CW	AEG-TFK 35 GHz vis-radar	
1978	SEL	35 GHz FM-CW	Japan Std control sensor	
	Nissan	24 GHz pulsed FM		
1980	Toyota Fujitsu	50 GHz FM-CW		Vehicular communication 60 GHz
1982	Nissan	60 GHz pulsed FM		
1984	Philips	94 GHz LC hybrid	AEG 60 GHz AVES	
1986	Un Lille	94 GHz Quasi-optical		
1988				MCCSV SEL / 60 GHz DACAR
1990				TST / 60 GHz AVES-COM

Table 2 Some of the millimetre wave application experiments that have been undertaken in the past

A NEW PASSIVE MICROWAVE SCANNER FOR

AIRBORNE MEASUREMENTS OF MARITIM OIL

POLLUTIONS

K. Gruener

DLR, Oberpfaffenhofen • Germany

A New Passive Microwave Linescanner
for Airborne Measurements of Maritime Oil Pollutions

K. Grüner, G. Kahlisch, H. Schreiber, P. Sliwinski
* B. Vowinkel

German Aerospace Research Establishment (DLR), Oberpfaffenhofen,FRG
* University of Cologne , Cologne,FRG

Abstract

A new passive microwave linescanner has been developed that will be used for airborne operational measurements of maritime oil pollutions. The system consists of two offset rotating parabolic mirrors and two radiometer sets, where each set contains three radiometers at 18.7, 36.5 and 89 GHz center frequency. A further 89 GHz radiometer is used for the measurement of the average radiometric sky temperature. The system is continuously calibrated by the use of a "hot load" at ambient temperature and a peltier cooled "cold load". A computer system allows online data reduction for the estimation of the amount of oil on the sea surface.

I. Introduction

Pollution of the sea has increased up to levels which can result in severe risks for the marine ecosystem. High loads of oil have been observed in many coastal waters. In addition to the procurement of "oil combating ships", this situation has led to the demand for airborne surveying methods for providing helpful data for oil spill clean-up operations following accidental discharges, and to monitor the legal aspects of permanent contributions to maritime oil pollutions.

In Europe, aircraft equipped with remote sensing instruments have been provided in various countries in the period 1983-1988 for a regular survey of coastal waters and, in view of a sensor system of second generation, for the purpose of basic research. Since 1986, for example, two Dornier Do28 aircrafts equipped with an IR/UV-Linescanner, a Side Looking Airborne Radar (SLAR) and a scanning 35GHz-Radiometer have been flown in the German-responsibility areas of the North Sea and the Baltic Sea, aiming at a continuous surveillance of maritime pollution and an improved guidance of oil combatting operations. In 1983,1985 and 1988, under the sponsorship of the European Community, a series of international airborne exercises took place in the North Sea (Archimedes Campaigns [1-3]). Different institutions

operated optical and active microwave sensors in parallel to a multitude of microwave radiometers (linescanners and profilers). The general aim of these exercises was to establish specifications for an improved operational sensor package for quantitative registration of oil pollution at sea under nearly all-weather conditions.

In 1989, the German Ministry of Traffic decided, to realize a new maritime surveillance system on board a modern Do228 aircraft [4]. In addition to improved sensors utilized up to this time in the first generation aircraft, the new system will include newly developed sensors, a Laser Fluorosensor and a multiple frequency microwave radiometer which allow a more thorough analysis of oil spills. This includes a quantification of the spilled volume over a range covering small discharges up to accidental events, and the classification of oiltypes.

II. The New Radiometer

A. Requirements

The Archimedes exercises offered a unique chance to study radiometer outputs in the 5GHz to 90GHz range for different types of oil and oilwater emulsions of varying thickness (up to several millimeter) during a variety of realistic environmental conditions. A series of findings resulted which, together with some operational restraints are now forming the basis for the development of the new radiometer. In detail, basic requirements for a radiometer of second generation are:

- simultaneous measurements at several frequencies to avoid ambiguity problems during the determination of layer thickness and, above all, the oil volume; an error of volume estimation smaller than 50% is tolerated,

- selection of frequency bands with reference to possible interferences due to man-made noise and registration of oillayer thickness in the $50\mu m$ to 3mm range,

- simultaneous registration of the sky radiation at least at the highest frequency band which is the most sensitive one with respect to variation of weather conditions,

- a scan angle as large as possible to realize wide scan widths,

- a spatial resolution as high as possible under the proviso of the maximum tolerable dimensions for the radiometer package (length:88cm, width:50cm),

- consideration of typical flight altitudes of 300m to 1000m and aircraft velocities near 70m/s.

B. Frequency Ranges

All requirements, listed above, have been considered during the development of the new radiometer. The construction of the linescanner was achieved with a compact configuration that realizes simultaneous imaging at 18.7, 36.5 and 89GHz. The frequency bands selected are reserved for passive measurements. During the Archimedes campaigns, measurements at 17GHz regularly resulted in the best volume estimations for thicker layers. A frequency around 35GHz seems to be the best compromise with respect to geometrical resolution and bad weather capability. 90GHz images are highly resolved and comparable in quality to IR images (Fig. 1). A sensitivity down to a layer thickness of 50μm exists. Oil-in-water emulsions and white caps (false alarm) can be determined qualitatively well in this frequency range.

In general, for optimum data evaluation, the sky radiation which influences the linescanner output should also be constantly mapped with a separate system in all frequency bands. But for simplification and cost reduction only one further 89GHz radiometer with a fixed sector-horn antenna has been designed for the sky radiation measurement. Its output delivers a mean-value of the relevant sky radiation which is the optimum basis for numerical estimation of the corresponding values at 36.5 and 18.7 GHz. The errors resulting from that procedure are expected to be smaller than 50%.

C. The Scanning Principle

Scanning happens with the aid of two parabolic mirrors installed in a continuously rotating cylinder (Fig.2). Beyond both front sides of the cylinder, identical three-channel-receivers are placed which are coupled to the parabolic mirrors via plane reflectors and through holes at the front sides. This scanning principle should be particularly advantageous with respect to operational use. It avoids oscillating movements of the main reflectors, rotational switches and guarantees redundance in the case of a failure of receiver channels. Further it supports a simple method for continuous calibration. A swath width of 476m is achieved for an altitude of 308m (1000ft).

Fig. 1. Example for the representation of passive microwave oil spill measurements. Oil slick taken at 90GHz during night time together with volume estimations (liter); altitude 4000 ft, scan width 440 m, maximum variation of brightness temperature 21 K.
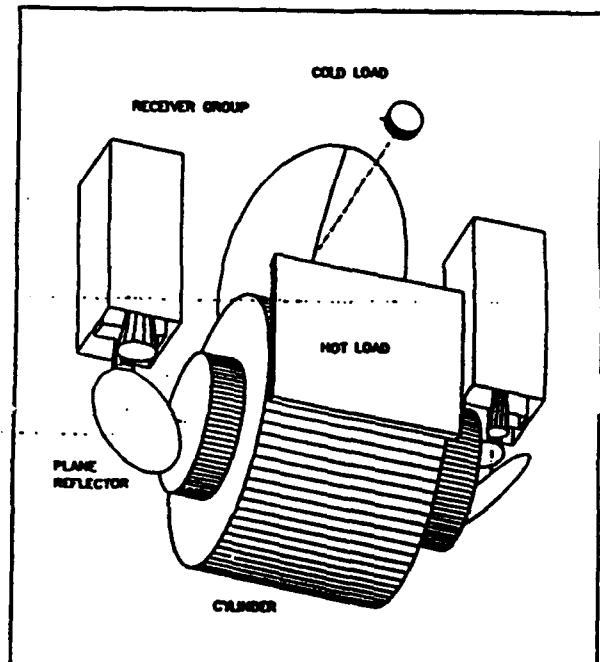
Fig. 2. Perspective representation of the MWR configuration developed. Two parabolic reflectors in a rotating cylinder and two receiver groups realize two scanlines per revolution. Thermal hot/cold sources are used for a continuous calibration.

D. The Calibration Principle

The calibration of the radiometer output signals is done by using a hot load at ambient temperature and a peltier-cooled cold load. The hot load consists of a sheet of absorbing material that is mounted on the inner wall of the radiometer main rack. As the effective area of the cold load is much smaller, a further mirror is used for focussing the beams of the rotating antennas onto the cold load. During each revolution of the cylinder a complete series of hot/cold measurements for all frequency channels is accomplished.

A cross sectional view of the cold load is shown in Fig. 3. The absorber is cooled by a 3-stage Peltier element that has a power consumption of 45W, this results in a temperature difference of 51K with reference to the case temperature. The core of the cooled absorber is made of aluminum in order to get a nearly constant temperature distribution up to the tips of the pyramids. The surface is covered with a layer of absorbing material. The reflecting inner wall of the load case has a conical shape resulting in an increased effective area of the cold load. This is important for the 18GHz channels which have the largest beam waist that is also somewhat off axis. The space between the inner wall of the case and the cold load is filled with small balls of expanded polysterene in order to prevent convection of the surrounding air and condensation of water. The overall efficiency of the cold load has been determined with an external cold load consisting of a large sheet of absorbing material cooled with liquid nitrogen.
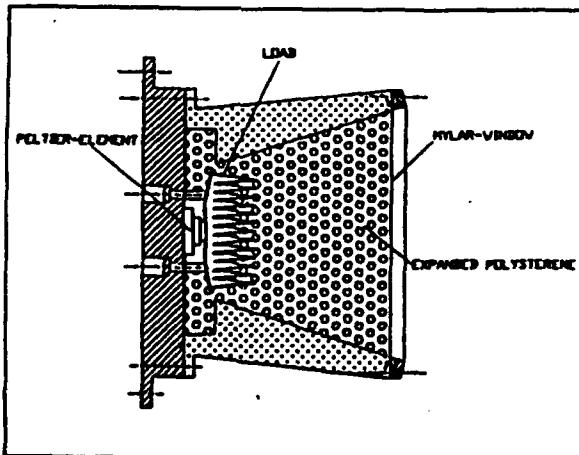
Fig. 3. Principal configuration of the cold load.

The sky-radiometer is calibrated by the same principle. However here we are using waveguide ferrite switches to switch the radiometer input to the loads. The cold load consists of a waveguide absorber that is also cooled by a 3-stage Peltier element.

E. Receiver Front- and Back-End Characteristics

To realize optimum radiometric resolution total power and heterodyne principles are applied to all channels. Balanced mixers are used throughout. They show the lowest noise figures which are, presently, obtainable with commercial devices at room temperature. The receiver housings shown in Fig.2 only contain the single front-end modules which can be simply removed like a drawer for better access during repair. The adjustment of a single radiometer channel for calibration happens by the use of a variable offset amplifier which is placed in a seperate back-end housing. Low pass filtering and A/D conversion follow after low frequency amplification.

The whole radiometer system is completely computer controlled. Not only calibration is done automatically but also a system configuration on the basis of a priority list (after switching on the radiometer). Further, during the measurements, a numerical correction of the antenna beam squint angels which result from the displacement of the feedhorns is realized. Digital processing, in addition, allows a continuous observation of the operational conditions of the system by monitoring and storing data such as the physical temperatures of the housings and calibration loads, the noise figures of the front-ends, and the gain and offset voltages of the variable amplifiers. Finally, in the case of oil spill measurements, layer-thickness and volume is directly processed from the received signals.

Different spatial resolutions are obtained for the single frequency bands by optimum utilization of the antenna apertures especially at 89GHz. However an identical resolution which would be important for an automatic unambiguous estimation of the volume, could be achieved numerically.

In table 1, the most important radiometer characteristics are listed.

III. First Airborne Measurements

After integration and installation into the aircraft, the new system has been flown for the first time in November 1991 in the Oberpfaffenhofen area. A special quick-look system had been developed for image representation, to qualify the radiometer. The first results showed that the concept of the new system works very well and that the resolutions meet the specified values (Fig. 4). On the other hand, a deeper analysis revealed some spuriuos interferences, among other things, due to local-oscillator radiation in the 18.7 and 36.5GHz ranges. These problems arise from the absence of unilines at the receiver inputs which normally results in higher sensitivities for well adjusted devices. In view of the first measurements above oilpolluted sea that are scheduled for April 1992, additional filtering in the 18.7GHz channels will be a short term solution. It is planned, however, to replace the het-

Tab. 1. Microwave Radiometer Characteristics

Antenna

Sensor:	linescanner	sky-radiometer
type:	rotating offset parabolic reflector	sector horn
angular resolution:	$0.8°$ / $2.0°$ / $4.0°$	$8°$ • $80°$
scan-with utilized:	$76°$	./.
scan-frequency:	10 revolutions/sec = 20 lines/sec	./.

Receiver front-ends

channel	type	center freq.	i.f.	noise temp.
1, 1'	heterodyne DSB	89.0 GHz	1-3 GHz	590 K
2, 2'	heterodyne DSB	36.5 GHz	.1-.5 GHz	565 K
3, 3'	heterodyne SSB	18.7 GHz	.6-.8 GHz	550 K
sky	heterodyne DSB	89.0 GHz	.1-1 GHz	590 K

erodyne principle by direct receiver principle, at an appropriate time, by using HEMT amplifiers in the 18.7 and 36.5GHz ranges.

Fig. 4. First airborne measurements with the new radiometer. 89GHz (left) and 36.5GHz (right) quick-look representation of a section with different types of vegetation or soils and water surfaces (black colour, storage lake). Demonstration of resolution capability and brightness temperature range awaited for polluted sea surfaces. Spurious interferences by external electronics. Altitude 3000ft.

References

[1] Gillot, R.H., Toselli, F.,
"The Archimedes 1 Experiment", Commission of the European Communities, Series: Environment and Quality of Life. EUR 10216 EN, 228 pp. Luxembourg 1985.

[2] Gillot, R.H.,
"The Archimedes 2 Experiment", Commission of the European Communities, Series: Environment and Quality of Life. EUR 11249 EN, 238 pp. Luxembourg 1987.

[3] Bekkering, J.A.,
"The Archimedes 2A Experiment", Commission of the European Communities, Series: Environment and Quality of Life. EUR 12674 EN, 83 pp. Luxembourg 1990.

[4] Grüner, K., Reuter, R., Smid, H.,
"A New Sensor System for Airborne Measurements of Maritime Pollution and of Hydrographic Parameters", GeoJournal 24.1 pp 103-117, 1991 (May), Kluwer Academic Publishers.

MMIC TECHNOLOGY FOR COMMUNICATION SYSTEMS

M. Aikawa

NTT • Japan

MMIC Technology for Communication Systems

April 22, 1992

Masayoshi Aikawa, Nobuaki Imai

NTT Radio Communication Systems Laboratories
1-2356 Yokosuka-shi, Kanagawa, 238-03 Japan

OUTLINE

MMICs are potentially very suited to high frequency bands
(microwave band to millimeter-wave band), due to

(1) mass productivity with high repeatability and excellent
    accuracy in the fabrication process,

and

(2) low cost potentiality.

TECHNICAL PROBLEMS
IN APPLYING MMICs TO COMMUNICATION SYSTEMS

(1) High cost at present
    because of the small market size and specialized needs,

(2) Circuit loss
    because of the small chip size

    ⟺ high Q resonator, filter, power efficiency, NF, etc.

(3) Lack of MMIC attractiveness for systems to which MMICs
    are applied

ADVANTAGES OF UNIPLANAR MMICs

- HIGH DESIGN FLEXIBILITY
- NO VIA-HOLES FOR GROUNDING NO POLISHING OF SUBSTRATE
- NARROW LINE-WIDTH
- SHORTER LINE SPACING
- COPLANAR CIRCUIT CONFIGURATION

- BALANCED/UNBALANCED LINE
- SMALL PARASITIC REACTANCE
- SMALL CHIP SIZE
- HIGH CHIP YIELD
- SIMPLE ASSEMBLY
- SIMPLE AND ACCURATE TESTING BY ON-WAFER MEASUREMENT

- HIGH PERFORMANCE
- LOW COST
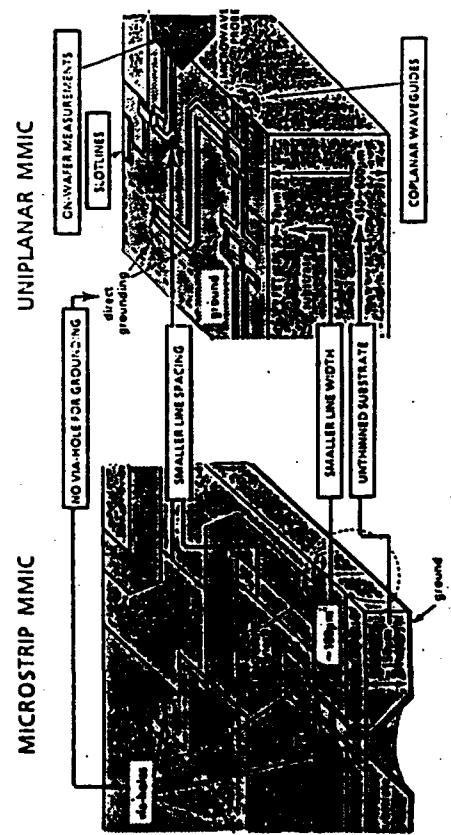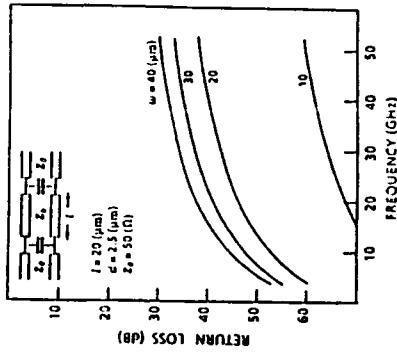
Characteristics of Air-Bridge

TECHNICAL APPROACHES TO OVERCOME THE PROBLEMS

(PROBLEM)   (APPROACH)

- Uniplanar MMIC
- Line unified FET (LUFET)
- Multi-layer MMIC
- Highly integrated MMIC with high yield
- Compounded MMIC technology
- Attractive performance unique to MMICs

- High cost
- Circuit loss
- Lack of attractiveness

MICROSTRIP MMIC   UNIPLANAR MMIC

- NO VIA-HOLE FOR GROUNDING
- SMALLER LINE SPACING
- SMALLER LINE WIDTH
- UNTHINNED SUBSTRATE
- ON-WAFER MEASUREMENTS
- SLOTLINES
- COPLANAR WAVEGUIDES
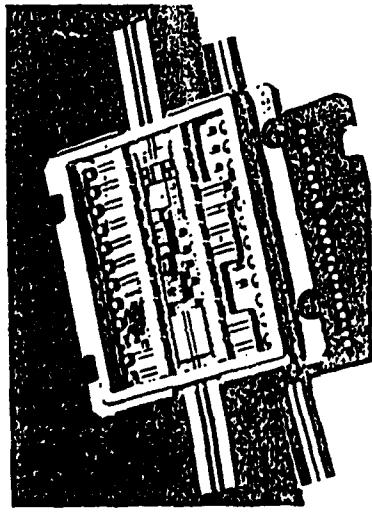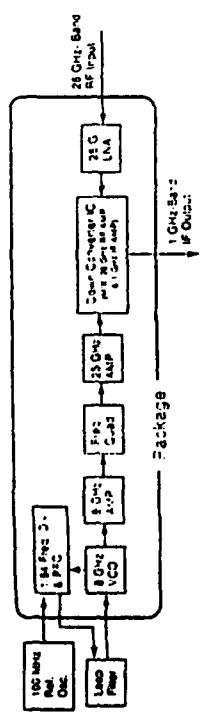
LINE-UNIFIED FET (LUFET)

Features:

(1) Super-compact functional circuits with transistor size,

(2) Remarkably wide frequency performance.

(a) Coplanar waveguide-to-slotline transition

(b) Power Combiner

(c) Power Divider
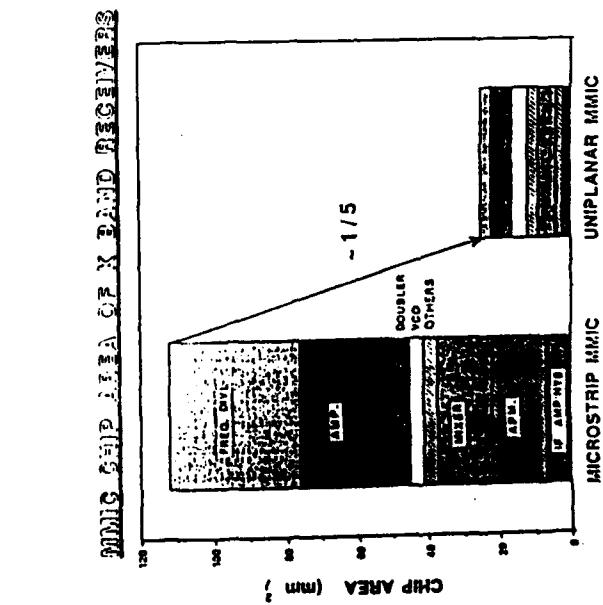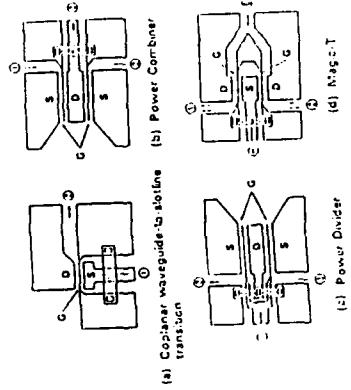
(d) Magic-T

Photograph of Magic T LUFET

26 GHz-Band MMIC RECEIVER MODULE

26 GHz Band RF Input

1 GHz-Band IF Output

Package

Package size (except flange) 21mm x 17mm x 3mm

MMIC CHIP AREA OF X-BAND RECEIVERS

~1/5

MICROSTRIP MMIC   UNIPLANAR MMIC

CHIP AREA (mm²)
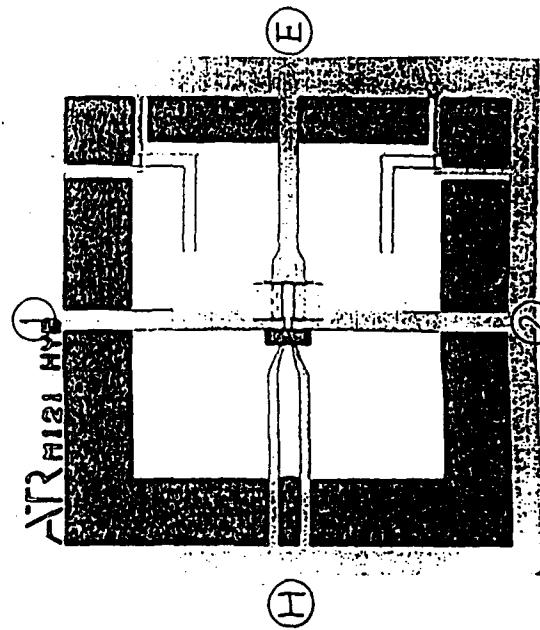
MINIATURE TRANSMISSION LINES FABRICATED USING A THIN DIELECTRIC FILM

(a) THIN FILM MICROSTIP (TFMS) LINE

(b) INVERTED E TFMS LINE

(c) QUASI-COPLANAR WAVEGUIDE (Q-CPW)

(d) INVERTED Q-CPW

Labels in diagrams: GND, Metal Strip, Dielectric Layer, GaAs Sub., W, H

LOSS CHARACTERISTICS OF TFMS LINE

P-CVD SiON:
Z0=50 Ω, H=9 µm, W=15 µm

Polyimide:
Z0=50 Ω, H=10 µm, W=22 µm

Line Length: 1.5mm

INSERTION LOSS (dB)

Polyimide

P-CVD SiON

FREQUENCY (GHz)

LOSS (dB/mm)

f=20GHz
f=10GHz
f=2GHz

t=1µm

t=H/3

SiON
ε_r =5.0

GaAs Sub.

FILM THICKNESS t (µm)

Line-Unified FET MMIC Magic-T Performance

ISOLATION (dB)

FREQUENCY (GHz)

COUPLING LOSS

RETURN LOSS

FREQUENCY (GHz)

MULTI-LAYER MMIC

Purposes are as follows:

(1) Substantially reducing MMIC chip size,

(2) Making high design flexibility for advanced MMICs possible,

(3) Achieving highly integrated MMICs.

Labels: Thin Film Microstrip Lines, Slotline, Coplanar Waveguide, Third Layer, Second Layer, First Layer, Substrate, FET, ground

26 GHz-Band Full MMIC T/R Breadboard Model

Receiver

Transmitter

30GHz-BAND MMIC RECEIVER FOR ETS-VI

A 12GHZ, 4-PORT WILKINSON DIVIDER MMIC USING VERTICAL CONNECTIONS

SIZE: 1.2 x 1.0 mm²

50Ω

100 Ω

Vertical Connections

TFMS Lines
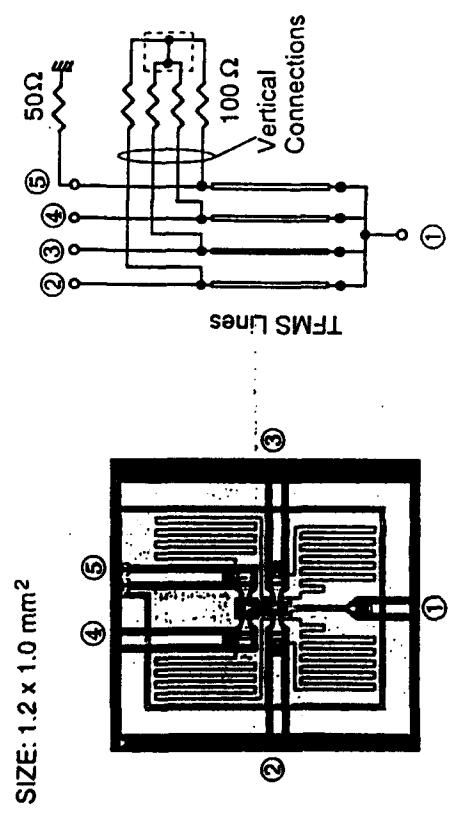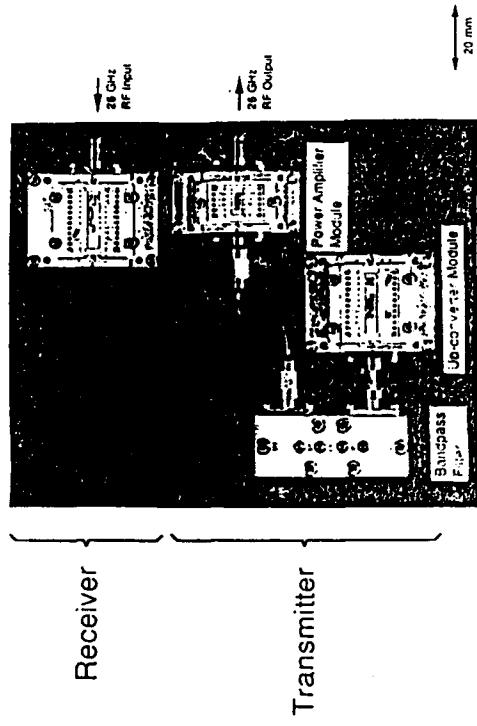
4/5/6 GHz-Band MMIC Transmitter & Receiver for 16QAM Digital Microwave System
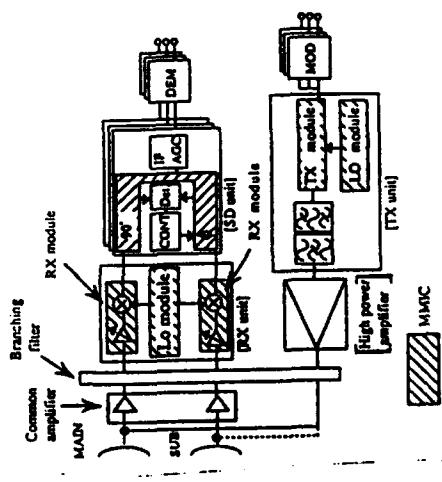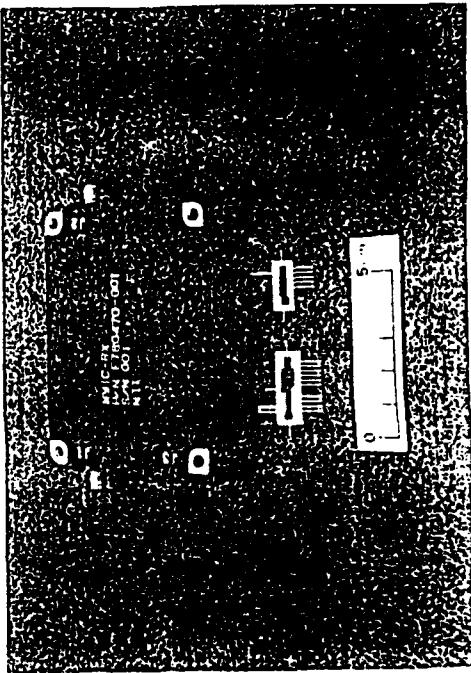
Prospects for MMIC Applications in the Communication Field

MMIC MARKET SIZE

DIFFUSION
PERIOD

WIRELESS LAN

PERSONAL COM.

SUBSCRIBER RADIO

MOBILE COM.

SATELLITE COM.

MICROWAVE TRANS.

YEAR

Now is the critical point for taking off.

(1) MMICs are now at the critical point for taking off.

(2) The initial application was satellite communication systems.

(3) Presently, digital microwave transmission systems are boosting the popularity of MMICs.

(4) The principal accelerators will be mobile communications, subscriber radio, wireless LANs etc..

CONCLUSION

1. Uniplanar MMICs and LUFET MMICs as well as highly-integrated MMICs are remarkably effective for the cost problem, and multi-layer MMICs are also very promising for the next generation.

2. In addition, the uniplanar-, LUFET- and multi-layer MMICs are very suited to high-frequency-band applications.

3. Performances / functions unique to MMICs and Compounded MMIC technology should be studied and developed more eagerly in order to overcome the disadvantages of MMICs, and to promote wider, more active MMIC applications in the communication field.

NEW INTEGRATED SUBSYSTEMS FOR MILLIMETER - WAVE APPLICATIONS

R. Funck

Dassault Electronique • France

NEW INTEGRATED SUBSYSTEMS FOR MILLIMETER-WAVES APPLICATIONS

Ronald FUNCK

Dassault Electronique
55 Quai Marcel Dassault
92214 Saint-Cloud, France

INTRODUCTION

The technology of millimeter wave circuits has been developing for many years and is now available for the realization of complex subassemblies to be used in various applications. The classical technologies are waveguide technology, finline or E-plane circuits, microstrip circuits, and are often used jointly in the same subsystem, depending on the performance and cost objectives.

Historically, the first millimeter wave applications have been for military (radars, sensors, seekers) and space (radiometers, communications) purposes. They have allowed, after the use of wave guide components, the development of planar and quasi-planar hybrid integration technologies (microstrip, fin-line or E-plane circuits). We can observe that, today, the development of new integrated subassemblies using these basic technologies is still continuing.

The emergence of new applications, especially civilian, and the search for cost reductions, have favoured the appearance of new principles of integration, among which the most promising are quasi-optical techniques.

Beside the development of hybrid circuits, much work is ongoing in the field of monolithic millimeter wave circuits development. Two types of technologies are concerned : Microwave Monolithic Integrated Circuits on GaAs (MMICs) and Silicon Monolithic Millimeter Wave Integrated Circuits (SIMMwICs), silicon having very interesting properties at such high frequencies.

This paper will first describe the new developments of millimeter wave subsystems for specific applications. After that, it will be focussing on more general or multi-applications developments.

RADARS AND SENSORS

The integrated subassemblies developed for these applications generally use a combination of wave guide, E-plane and microstrip circuits.

The finline or E-plane technique was one of the most promising hybrid integration techniques and research was being carried out worldwide [1], as this technology had emerged as a front-runner. The finline structure, which is a planar printed circuit enclosed in a wave guide like split-block housing, has relatively low loss, and the conductor patterns are on circuit boards, which can be made reproducible using conventional photolithographic techniques. Entire systems were designed and realized using finline technology. We will describe some recent developments made at Telefunken System Technik of Ulm, Germany [2,3,4].

The block diagram of a polarization agile 94 GHz monopulse radar frontend for interrupted FM-CW operation with 600 MHz bandwidth [2] is given in fig. 1. The transmitter part of the radar frontend is formed by a 500 mW pulsed IMPATT oscillator, which is injection synchronized by a bias tuned CW Gunn oscillator. A high isolation SPST switch with 60 dB isolation and 1.6 dB insertion loss is inserted between Gunn and IMPATT oscillators in order to ensure maximum receiver sensitivity during pulse intervals. Transmit-receive diplexing is carried out by E-plane circulators, integrated in common housings with finline balanced receiver mixers and IF amplifiers. The technology used for the design of the frontend is a combination of E-plane and integrated waveguide milled block technology, enabling straight forward design and manufacturing of integrated subsystems in the frequency range 30–150 GHz. The frontend overall design and its separation into modules suitables for easy production and testing is shown in fig.2. The global frontend performances are : 270 mW output power, 4 dB receiver conversion gain.

The integrated milled block technology has been used for the design of an active passive sensor frontend in the 90 GHz range [4]. This combined radar/radiometer sensor, the block-diagram of which is given in fig. 3 and its photography in fig. 4, has been developed for a low cost application. The Gunn oscillator provides transmitting power (30 mW) as well as local oscillator power for the mixer. The received signal is converted down to video frequency which can, after being amplified, be separated by filtering into radar and radiometer channels. The entire RF unit is very compact and measures only 20 mm x 20 mm x 23 mm, and has a mass of 16 g. A number of complete subsystems have been successfully tested at accelerations reaching 16000 g.
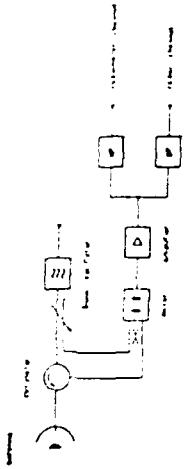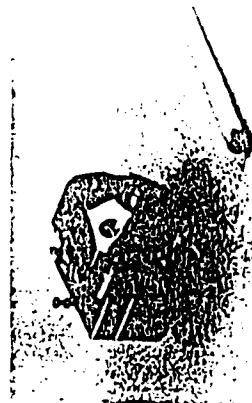
These sensors were originally developed for military purposes. Today they are under consideration for automotive radar applications, due to their potential low cost.

Microstrip technology has been used by GEC, U.K. for the production of millimeter wave circuits for more than 10 years. Its development progressed upward in frequency, with 35 GHz components

Fig 2 : Radar frontend components [2]

Fig 1 : Block diagram of interrupted FM-ICW radar frontend with polarization agility [2]

Fig 3 : Block diagram of a combined passive/active sensor frontend [4]

Fig 4 : Active/passive sensor [4]

being developed in the mid 1970s and 94 GHz components in the 1980s. Microstrip components are now available to operate up to 220 GHz [5] This technology uses hard substrates such as single crystal quartz or glass ceramics, which allows the insertion of ferrite pucks for iso-circulators and provide compact, rugged assemblies.

The fig 5 gives an example of a fully integrated 94 GHz FM-CW transceiver using this microstrip technology [6] which has been designed specifically for an anti-armour sensor requirement. The transceiver is realized in a volume of 1 cubic inch. A single quartz microstrip circuit contains all the millimeter wave components, except the VCO, which is in a waveguide cavity. The transmit power is approximately +8 dBm, and the system overall noise figure is less than 16 dB.
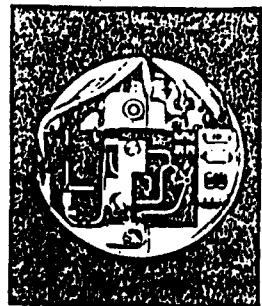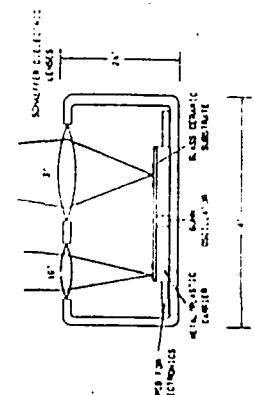
thin film microstrip circuit. A quasi optical concept is used for the antenna : patch antennas on the microstrip circuit are used to illuminate a dielectric lens. By illuminating the lens from different patches, a beam steering effect can to achieved. The dielectric lens forms an hermetic seal and can be produced as part of the housing.

Microstrip technology has also been used for a millimetric sensor for stand-off weapons realized by Marconi Defence Systems, U.K. [8]. The received signal is fed from the antenna to the receiver via a quasi-optical feed. This is capable of providing the necessary R.F. input to the receiver so that co- and cross-polar sum, elevation and azimuth error channel data may be simply extracted. The R.F. receiver uses quartz microstrip techniques which enables the integration of all the passive microwave components (mixers, isolators, couplers, etc.) on a single circuit. The LO uses a Gunn diode in a waveguide cavity (10 to 15 mW of output power). The transmitter uses an IMPATT in a waveguide cavity (300 mW) which is injection locked to a Gunn diode VCO (20 to 30 mV) swept in a linear fashion. The linearity of the sweep is controlled by the use of a closed loop system which maintains the frequency ramp at a given rate.

A very interesting concept has been described by the RSRE, U.K. [9]. This paper describes a staring, multibeam, focal-plane array technology and radar development which has evolved from research into lens-fed microwave and millimeter wave receivers with integral antennas. The key components are a high permitivity dielectric lens and a focal-plane array of receivers in intimate contact with the lens (Fig 8). The lens generates a multiplicity of simultaneous narrow beams, each with its own direction of lock, covering between them the total field of view. Each beam is focussed to a unique position on the lens focal-plane. Combining this with an array of receivers in the focal-plane creates an array of high gain receivers which collectively stare across a wide field of view, with no moving parts and without separate beam forming circuitry. Each receiver is built round a cross-dipole antenna, in which one dipole receives the signal and the other the local oscillator This provides a convenient means to drive a quad diode mixer (Fig 9). Receiver arrays of dipole coupled mixers replicated on a single high permitivity substrate have been realized a 35 GHz 4x4 monolithic silicon array and 5x5 hybrid arrays on thin-film have been realized Typical noise figure of receivers of this type were measured to be 8.5 dB including 2.0 dB IF noise A 35 GHz binodyne radar demonstrator has been built, incorporating an array of nine 5x5 thick-film receivers As can be seen this is a good example of a system combining different technological concepts : monolithic circuits on silicon and quasi-optical techniques

Fig 9 ...

COMMUNICATIONS

The main applications are for in-door communications or communications with mobiles at 60 GHz and for satellite communications. The technologies which are used are of the same type as previously described.

T.S.T. has designed a 2 1 W solid state transmitter for short range data communication at 60 GHz [10] composed of 3 stages of amplification. The first uses a reflection type Gunn diode amplifier (20 mW), the second an injection locked Gunn diode oscillator (150 mV) and the third, an injection locked

Fig 5 94 GHz FM-CW transceiver (microstrip)

Fig 6 Proposed structure of a lens transceiver [ ]

The same technology has been used to design a millimeter wave system for cruise control on automobiles [7].This system is operating at 77 GHz and comprises an FM-CW radar realized in thin film microstrip technology and a switched beam quasi optical antenna. Active cruise control enables the driver to engage cruise control when travelling at a 'driver determined' safe distance behind a vehicle. As the selected vehicle speeds up or slows down, your vehicle tracks this change by accelerating or decelerating. A proposed structure of the control unit is given in fig 6. The system (fig.7) is a basic FM-CW radar based on a varactor tuned waveguide Gunn oscillator using the radial disk geometry with extraction of the second harmonic (30 mV output power, 1 GHz tuning range, -85 dBc/Hz phase noise at 100 kHz off carrier). The received signal is mixed with the oscillator sampled with a 10 dB coupler to derive LO power. A further sampling coupler is used in a FM feedback loop which is used to linearize the ramped output of the oscillator : the 77 GHz output signal is downconverted in the 850-1150 MHz range with a subharmonic mixer driven by a 14 GHz DRO and applied to a 10 MHz frequency discriminator through a 1/100 frequency divider. All the microwave components are located on a single

Fig 7 Block diagram of the cruise control unit [7]

Fig 13 : 30 GHz MMIC Amplifier [19]

AUTOMOTIVE APPLICATIONS

Many programs related to automotive applications are now under study in Europe. In particular, there are european projects like DRIVE and EUREKA. The main applications are : automatic debiting systems, Doppler speedometers, vehicle-vehicle and vehicle-road communications, obstacle detection. These programs could lead to large microwave applications if two conditions are satisfied : low cost technology compatible with mass production, low size and low weight design to fulfil the implementation requirements.

The search for low cost and low volume at millimeter wave frequencies may force the engineer to simplify components and to combine functions whenever possible. The use of MMICs could be appropriate, but these components doesn't solve the problem of circuit integration.

Rather than scaling low frequency techniques to shorter wavelengths, a completely new approach is often needed. Quasi-optical components provide a solution to this problem.

An example of realization of a radar sensor for automobile cruise control, directly derived from radar developpements, has been described above [7].

T.S.T, Germany, has developed a very simple radar sensor for traffic data acquisition system named AVES [12]. This system obtains specific traffic information from appropriate Doppler frequency The chosen frequency of operation, 61.5 GHz, offers several avantages : the band is available for this application in Europe (ISM-band, Industrial, Scientific and Medical). Due to the high frequency , small antenna dimensions allow good focussing, short range application and economic reuse of frequencies due to high atmospheric absorption. The sensor is relatively small (150x150x150 mm^3), and can be mounted on existing sign posts and bridges. The schematic set-up of the mm-wave front end is presented in figure 14. Hybrid integration was implemented, combining finline and waveguide to reduce cost : this unit is composed of a cavity second-harmonic type Gunn oscillator, a waveguide filter and a finline mixer. Vehicle speed, vehicle length and the distance between vehicles is obtained by the AVES sensor. A future application of this type of sensor can be seen in vehicle-roadside communications.

Fig 14 : The AVES system [11]

GaAs IMPATT oscillator (2 W). The block-diagram of this amplifier is given in fig 10. The breadboard has been realized in waveguide technology.

The work done by Siemens Telecommunicazioni in Italy [11], concerns commercial inter-orbit and inter-satellite communication applications. In this sens, advanced studies are in progress in order to evaluate a possible experimental mission of a 59/64 GHz communication payload. Their work demonstrates that resistive mixers present a reliable and performing solution for mm-wave low noise mixers. The mixer which has been realized uses a cross-bar configuration which allows a clear separation of the various LO sideband harmonics, as well as image recovery. Tests have been made at 40 GHz with a 4 GHz IF. Insertion loss is about 3.2 dB in a 1 GHz bandwidth (fig. 11), showing clearly the performance improvement due to image frequency recovery. It is expected that at 65 GHz, the conversion loss would be 3.7 dB.

Fig 10 : Block diagram of the 2W transmitter [10]

Fig 11 : Image recovery 60 GHz mixer [11]

The European AIMS/ESPRIT programme, led by Thomson-CSF, France, adresses components up to subsystems level for applications between 20 and 30 GHz. System pulls are from short-hop land links and ultra-small aperture satellite terminals (USAT). A range of generic MMICs are being developed, covering switches, VCOs, solid-state amplifiers. Apart from the T/R switch, which makes use of MESFETs, all other circuits are based on the use of heterojunction transistors : HEMT, P-HEMT, HBT. These activities will be briefly described in a next paragraph. Within this project, Thomson-CSF/TCM is developping a 27.5-30 GHz synthetiser. The fig. 12 and 13 represent respectively the microwave part of the synthetiser and the 30 GHz MMIC amplifier chip. The main performances are : 10 dBm output power, less than -70 dBc/Hz phase noise at 100 kHz off carrier, and 27 MHz frequency steps. The subassembly consists in cascaded buffer, frequency divider, buffer, VCO, power divider, frequency multiplier, amplifier. At this step, the fonctionality has been proven with conventional thin-film hybrid technology to interconnect individual MMIC chips. Later on, compact versions will be designed together with comparing improvements gained from advanced MMIC technology.

Fig 12 : 30 GHz synthetiser for AIMS

A first example of the quasi-optical approach is given by the design of ALDER from the RSRE [3].

The University of Lille, France, presents an other circuit approach based on a quasi-optical planar integration [13] which suppresses the physical connections between functions, the transfer of energy being achieved either by electromagnetic radiation or by electromagnetic coupling. This allows the realization of low cost active antennas which can be coupled to Fresnel lenses to fulfill the various directivity requirements. The concept of this antenna is given in fig.15. The transmitter is composed of a circular patch fed by two terminal devices (Gunn or IMPATT diode) in a TM_{11} radiating mode. The insertion location of the diode is chosen to optimize its matching. Electronic frequency tuning is provided by a varactor which is implemented in parallel inside the circular patch cavity, the value of the coupling depending on the distance from the disc center. The receiver is realized by the means of a Schottky diode mixer integrated on the patch. The principle is to isolate a strip following the TM_{11} mode current lines, the receiver diode being connected in serie on this strip. The concept has been demonstrated at X-band. Some modifications have been made for mm-wave applications, in order to increase the size of the antenna to allow the use of packaged Gunn or IMPATT diodes. This is achieved by coupling a TM_{11} circular patch antenna to a TM_{12} annular patch antenna by means of a circular slot antenna as illustrated in fig.16. In addition, simple electronic beam forming can be obtained by proper perturbation of the electromagnetic fields : scanning of the beam can be achieved at +/- 10°. Two anticollision radar prototypes have been developed : a pulsed 60 GHz Gunn radar which gave excellent results for a 150 m range, and a 86 GHz FM-CW IMPATT radar which exhibited lower sensitivity due to the IMPATT diode noise.
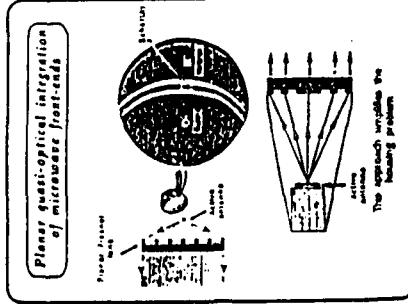
QUASI-OPTICAL

Quasi-optical techniques can also be used as an elegant way to improve the power handling by coupling many devices distributed over a surface several wavelengths in dimensions [14]. This technique is referred to as quasi-optical array or grid technique. This technique can be applied to oscillators, multipliers, phase shifters and mixers. The fig.17 illustrates the principle of a grid oscillator.

Theoretical modelling of grid power combiners is being made by the LEST, France [15].

Quasi-optical techniques are also well adapted to the realization of filters. The LEST is studying planar grating-reflector antennas [16] on a theoretical point of view. Dichroic filtering structures are being realized for radiometer applications by MATRA ESPACE, France, [17]. Such structures are transparent or reflective depending on the wavelength of the incident electromagnetic wave. A dichroic

Planar quasi-optical integration of microwave front-ends

Fig 14 : Integration of the emitting diode [13]

⊗ IMPATT or GUNN diode

Fig 15 : [13]

triplexer using resonant grid structures etched on a fused silica substrate has been realized demonstrating very low loss : insertion loss at 90 and 157 GHz is 0.2 and 0.3 dB, reflexion loss at 183 GHz is 0.2 dB. A different approach is used by the Institut National de Télécommunications and the Observatoire de Meudon, France, to be used for Météosat satellite [18]. The dichroic triplexer are realized by using perforated screens. The performances obtained are : 0.54 dB insertion loss at 183 GHz, 0.05 and 0.11 dB reflexion loss at 150 GHz and 110 GHz respectively.
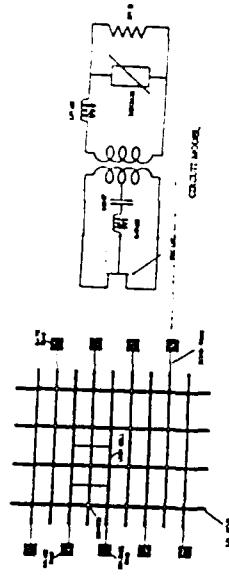
Fig 17 : Planar grid oscillator

MONOLITHIC INTEGRATION

Two technologies are used for the realization of monolithic integrated circuits and subsystems : Microwave Monolithic Integrated Circuits on GaAs (MMICs) and Silicon Monolithic Millimeter Wave Integrated Circuits (SIMMWICs).

MMICs

Three ESPRIT programs are concerned with MMICs developments for mm-wave applications.

AIMS adresses components and subsystems for applications between 20 and 30 GHz. The fig.13 shows a 27-40 GHz amplifier from Thomson, which is used in the synthetisor described above. The gain is 15 dB, noise figure 14dB and size 1.5 mm². This amplifier uses a 0.25 μm pseudomorphic technology [19]. The same laboratory has designed a 60 GHz LNA having a 12 dB gain and a 5 dB noise figure.

MONOFAST is focussed at 44 GHz. The overall aim of this project is to integrate unfamiliar technologies and new design tools into a high proven methodology for high frequency MMICs. An MMIC LNA at 44 GHz based on sub-micron MESFET and coplanar waveguide technology is being designed by Alcatel Espace, France [20] : a gain of 15 dB and a noise figure less than 4 dB have been simulated.

GIANTS is concerned with the development of InGaAs based transistors where one of the aims is to demonstrate a low noise MMIC amplifier at 60 GHz.

Philips Microwave Limeil, France, has developped several mm-wave MMICs using a pseudomorphic HEMT technology [21, 22]. A push-push oscillator using a 0.5 μm HEMT has been realized demonstrating a -3 dBm output power and a phase noise of -75 dBc/Hz at 1 MHz off carrier at 52 GHz. An image rejection mixer (fig.18) designed in the 52-60 GHz frequency range and using a 0.25 μm HEMT has a typical conversion loss of 14 dB and an image rejection higher than 30 dB for a 2-4 GHz IF frequency range.

In Germany, Telefunken System Technik is also engaged in an important mm-wave MMICs development program [23, 24, 25]. Using the integration of MESFET devices (Imax = 90 GHz) and Schottky diodes (Ft = 2300 GHz) on the same chip, mixers working at 35 GHz and 60 GHz have been realized. The 35 GHz mixer, integrating a 2-stage IF amplifier, has a 6 dB conversion gain with a 9 dB DSB noise figure. The 60 GHz receiver chip has a 2-stage IF amplifier : the conversion gain of this chip is 12.5 dB with an associated DSB noise figure of 9.5 dB. The fig. 19 represents the 60 GHz receiver chip with a chip size of 3x3.5 mm². The mixer itself consists of a Lange-coupler feeding two Schottky

dodes and LO and signal bandstop filters at the IF output. The diodes have a finger width of 0.3 μm. The low noise IF amp is employing two MESFETs with a gate length of 0.75 μm. Further developments have been made using a P-HEMT technology, which has proven to be superior to all other available monolithic technologies for both a reduced low noise and power circuits in the mm-wave frequency range. Quarter-micron HEMTs with a Fmax of 200 GHz have been developed and fabricated. Using these devices, a 2-stage low noise amplifier has been fabricated having a gain of 5 dB at 53 GHz.

Fig 18 94 GHz receiver mixer MMIC

Fig 19 94 GHz receiver chip [10]

SIMMWICs

The extension of transmission lines realized on high resistivity silicon (>2000 Ω cm/cm) is dominated by coplanar access around 40 GHz. On the other hand, silicon substrates allow the integration of two terminal devices such as IMPATT diodes, Schottky and PIN diodes, PN diodes. Therefore, silicon is also a good candidate in the design of monolithic integrated circuits in the millimeter wave range.

Monolithic integrated 94 GHz transmitters have been fabricated by Daimler Benz Research, Germany. The transmitter uses a two terminal sub-harmonic IMPATT diode as the active in a radiating structure. The resonator uses a high resistivity silicon substrate (fig 20). This subsystem is realized using the self-oscillating mixer principle. For simplicity of the layout, the resonator and the antenna are merged into one. The resonator is also a resonator. This model of antenna resonators needed, the characteristic of radiating antenna, coupling losses between antenna and free space are avoided. Antenna resonators are made as a length of transmission wavelength. The IMPATT diode is integrated in the center of the resonator as a length of the substrate wavelength. For biasing, the diode has small pads are provided as a quarter wavelength apart from the ends of the line dc resonator. For beam losses power supply are provided via a dc/dc which is provided up to now. 26 mW CW and 50 mW for short pulse operation for pulses duty cycle 0.5% have been obtained at 90 GHz. The spectral purity sensitivity, profile line is typically 50 dB/Hz at 100 kHz offset.

Fig 20 SIMMWIC 94 GHz transmitter [??]

Fig 21 SIMMWIC planar antenna diode [??]

Planar millimeter wave silicon detector circuits have been realized by the same laboratory [??]. Schottky barrier diodes have been monolithically integrated with planar antenna structures on high resistivity silicon. The Schottky barrier diodes were fabricated by MBE. Now series resistance (<5 ohms), ideality factor of less than 1.1 and cut-off frequency up to 1 THz have been achieved. The fig 21 gives the layout of two types of receiver which have been developed and fabricated. With the two antenna structure, detection sensitivities of 40.8 mV/(mW/cm²) without bias, and of 90 mV/(mW/cm²) with 280 mV bias, have been achieved at 94 GHz.

CONCLUSION

In reviewing the progress made in the last years in millimeter wave subsystems, we have seen that the search for low cost and large volume quantities for commercial applications such as communications, traffic control, collision avoidance, etc ... has favored the emergence of new principles of integration in parallel with the development of monolithic millimeter wave circuits.

Millimeter wave applications and technologies, which have been increased and pushed by military needs first, are now available for new commercial applications.

REFERENCES

[1] H.H. MEINEL, "Millimeter Wave Technology Advances Since 1985 and Future Trends", IEEE Trans Microwave Theory Tech, vol MTT-39, pp 759-766, May 1991

[2] L.P. SCHMIDT, et al "94 GHz FM/CW Micropulse Radar Sensor", Military Microwave 92, pp 181-186, July 1990

[3] W.H. CLIFP, "Radar and Radiometer Sensors on Millimeter wave", Benz art, custom, 8-11 (1989) H 4, pp 155-173

[4] W.H. CLIFP, "Design and Technology of a Monolithic Integrated Accelerator on band detector end Millmeter Gen Range", Milary Microwave 90, pp 103-114

[5] D.A. WILLIAMS, "Millimeter-Wave Components and subsystems B, Using Microstrip Technology", IEEE Trans Microwave Theory Tech, vol MTT-35, pp 711-774, May 1990

[6] P. LOWBRIDGE, "Low Cost 94 GHz FM-CW Radar for Precision Range Measurements", Military Microwaves 92, pp 311-319

[7] P. LOWBRIDGE, et al "Realisation of a Low Cost Millimeter Wave Cruise Control System for Automobile Applications", private comm

[8] M.R. NICHOLLS, "A Millimetric Sensor for Stand-off Weapons", Military Microwaves 92, pp 187-175

[9] C. J. ALDER, et al "Microwave and Millimeter-Wave Staring Array Technology", 1991 IEEE MTT-S Digest, pp 129-1252

[10] H. BARTH and M. PIRKL, "A 2 W Solid State Transmitter for Short Range Data Communication at 61 GHz", 1990 IEEE MTT-S Digest, pp 931-933

[11] M. PILON and G. CARPENTIERE, "60 GHz Down Converter for Satellite Low Noise Receiver", European Microwave Conf, pp 145-150

[12] W. STAMMLER, et al "MM Wave Sensor for Traffic Data Acquisition", Proc of Proc Int Microwave Symp (Sao Paulo, Brazil, 1989
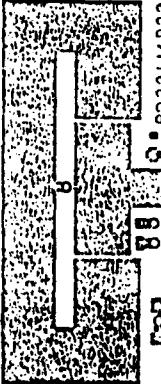
[13] P.A.ROLLAND, et al.,"Quasi optical Technology for MM-Wave Car Electronics", *Workshop 21st EuMC 91*

[14] E.KOLLBERG,"New Solid State Devices and Circuits For Millimeter Wave Applications", *European Microwave Conf. 1991*, pp.36-54

[15] A.TCHAMENI,"Surfaces sélectives en fréquence", *Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp.57-60

[16] H.CAM,"Addition de puissance en ondes millimétriques",*Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp.65-68

[17] J.M.GOUTOULE,"Dichroïque millimétrique à grande incidence réalisé en photogravure", *Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp.69

[18] C.LETROU,"Conception de filtres dichroïques pour le millimétrique et le submillimétrique", *Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp.84-87

[19] P.BOURNE,"Amplificateurs faible bruit à 35 et 60 GHz", *Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp. 39-42

[20] C.TRONCHE ,"Fonctions millimétriques à 30 GHz, 44 GHz,et 90 GHz pour applications spatiales", *Journées d'études sur les composants millimétriques,Carry-le Rouet (France)*, pp.1-16

[21] P.GAMAND,"L'intégration monolithique dans le domaine millimétrique",*Journées d'études sur les composants millimétriques, Carry-le Rouet (France)*, pp.43-46

[22] P.GAMAND,"Développement des MMICs en gamme millimétrique", *2èmes Journées d'étude Micro-ondes et Espace, Toulouse, France*, Jan 1992, pp.153-156

[23] B.ADELSECK, A.CLOQUHOUN,"MMIC Receiver Modules for Frequencies up to 100 GHz",*Military Microwave 90*, pp.305-310

[24] B.ADELSECK, et al.,"Monolithic 60 GHz Amplifier using Low Noise pseudomorphic HEMTs", *European Microwave Conf. 1991*, pp.341-345

[25] B.HUDER, et al.,"Recent MMIC Developments and Concepts Leading to Advanced Microwave and Millmeter Wave Phased Array Antennas", ESA Workshop on Advanced Beamforming Networks for Space Applications, nov.91

[26] J.BUECHLER, et al.,"Coplanar Monolithic Silicon IMPATT Transmitter", *European Microwave Conf. 1991*, pp.352-356

[27] K.M.STROHM, et al.,"Planar 100 GHz Silicon Detector Circuits", *ESSDERC'91*, Lausanne, sept. 1991,

NEW MILLIMETER - WAVE DEVICES REALIZATION

AND PERFORMANCE

D. Pons

Thomson - CSF LCR • France

FIELD - EFFECT TRANSISTORS FOR THE MILLIMETER - WAVE RANGE: PHYSICAL ANALYSIS, MODELING AND EXPECTED PERFORMANCE

G. Salmer

Université de Lille • France

iemn

OUTLINE

● **PHYSICAL DEVICE MODELS**

● **LOW NOISE DEVICES**

- Noise simulation
- MESFETs
- MODFETs on GaAs Substrate
- MODFETs on InP Substrate
- Discussion

● **POWER DEVICES**

- Power requirements
- Breakdown phenomena
  - Power FETs.

C.H.S.

---

iemn

**FIELD EFFECT TRANSISTORS
FOR THE MILLIMETER WAVE RANGE**

**Physical Analysis - Modelling
and Expected Performance**

G. SALMER

I.E.M.N. - U.M.R. CNRS 9929

Département Hyperfréquences
et Semiconducteurs
Université des Sciences et Techniques de Lille
Villeneuve d'Ascq - France.

With the help of
- E. CONSTANT
- Y. CROSNIER
- A. CAPPY
- R. FAUQUEMBERGUE
- P. BOURREL

C.H.S.

PHYSICAL MODELLING OF F.E.T. : METHODOLOGY
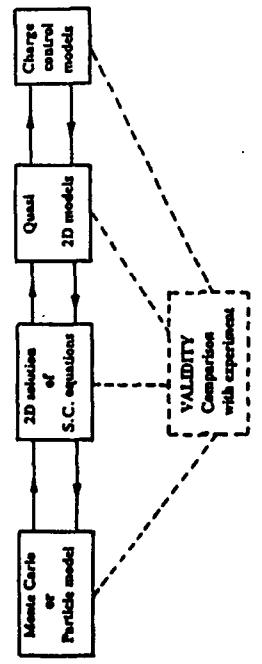
MONTE CARLO OR PARTICLE MODELS

Principle

The stochastical motion of the carriers is studied simultaneously:

- in the 3D k space : for each time step Δt, are considered

- the effect of the electric field : $\vec{\Delta k} = \frac{q}{h} \vec{E} \Delta t$.
- the effect of scattering mechanisms.

- in the 2D r space : that gives the two dimensional description of

- electric fields
- carriers concentrations
- carrier velocities

from which can be deduced   - external currents and voltages
                                              - equivalent circuits

Remark : other possibilities : Direct solution of the Boltzmann transport equation.

MONTE CARLO OR PARTICLE MODELS

- **Advantages :**
  - S.C. Equations automatically solved
  - All { physical effects } can be included
  - { transport properties }
  - Determination of all device properties.

- **Requirements :**
  - Accurate description of energy band diagram
  - Knowledge of various scattering mechanisms

- **Problems :**
  - Bad knowledge of several effects : screening, ionization
  - Transport properties unknown for several ternary or quaternary compounds
  - Introduction of quantum effects

- **Drawback :**
  - Very long computational time
  - Powerful computers
  - Difficult to study dynamic regimes.

MONTE CARLO SIMULATION
OF A AlInAs/GaInAs MODFET :

PARAMETRICAL STUDY : INFLUENCE OF AlInAs LAYER CHARACTERISTICS
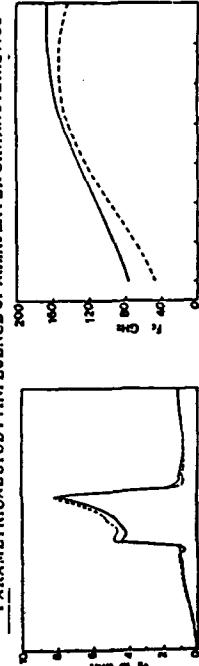
Longitudinal velocity

AlInAs layer :   - $N_D = 4\ 10^{18}$
                      - $N_D = 2\ 10^{18}$

Internal cut-off frequency :

a = 75 Å   ———
a = 125 Å   - - - - -

from Bourrel and Fauquembergue, 1991

EXEMPLE OF QUASI 2D MODEL
HELENA SOFTWARE

Device Structure
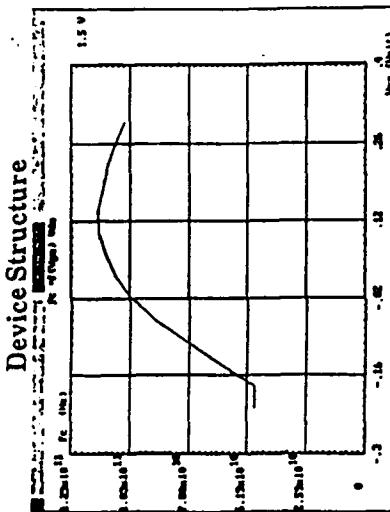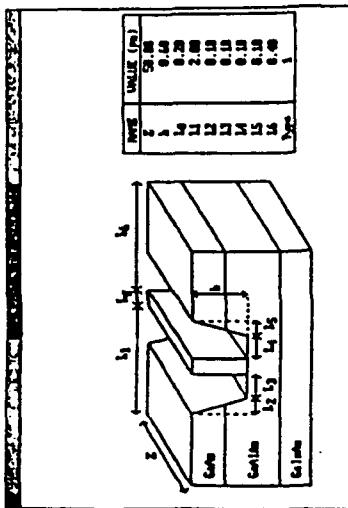
Current gain cut-off frequencies predicted

H. Happy, A. Cappy, G. Dambrine, 1991.

C.H.S.

INFLUENCE OF GATE TO DRAIN
CAP LAYER SPACING R2

R$_2$ μm
.05
.15
.30

R$_2$ μm
.30
.15
.05

Output Conductance Maximum oscillation frequency
for L$_g$ = 0,3 μ   R$_1$ = 0,05 μm

Following Shawki and Salmer (1990)

QUASI TWO DIMENSIONAL MODELS : ONE DIMENSIONAL SOLUTION
OF SEMICONDUCTOR EQUATIONS

● **Main assumptions :**

- Equipotential lines perpendicular to the source to drain axis
→ one dimensional treatment
- Depletion regions are fully depleted.

● **Main advantages :**

- Computational effort is strongly reduced
- Main physical effects can be account for : overshoot, surface, buffer injection, ionisation...
- Parametrical studies are easy
- Large signal behaviour and noise properties can be deduced.

● **Problems :**

- Difficulty to treat very complicated structures or two dimensional effects
- Validity for very short gate devices.

C.H.S.

2D SOLUTIONS OF SEMICONDUCTOR EQUATIONS - TYPICAL RESULTS : MESFET SIMULATION

Electron concentration

D.L.T. occupancy

Hole Density

Effect of substrate electron injection on acceptor deep trap level occupancy.

From Ghione et al, 1990.

C.H.S.

TWO DIMENSIONAL SOLUTIONS OF SEMICONDUCTOR EQUATIONS

· **Principle:**   Two dimensional solution of:

- Equations deduced from B.T.E.: continuity, momentum relaxation, energy relaxation.
- Poisson equation
        By taking into account
- Physical effects and material properties : trapping in the substrate, surface potential, ionization effects
- Parameter values deduced from Monte Carlo simulation : τ, μ

· **Remarks:**

- Various kinds of models corresponding to various degrees of assumptions : single electron gas, drift diffusion models.
- Possibility to account for quantum effects by partially coupling with Schrödinger equations
- Finite difference methods are generally used.

C.H.S.

TWO DIMENSIONAL SOLUTIONS OF SEMICONDUCTOR EQUATIONS

· **Advantages:**

- The most complete equations allow to account for non steady state transport phenomena and other physical effects (trapping, surface, ionization...).
- Dynamic characteristics, transient response and microwave performance can be deduced
- Parametrical studies can be made

· **Problems:**

- Validity of semiconductor equations or simplified assumptions (drift diffusion for submicrometer gates)
- Inclusion of quantum effects
- Treatment of · Schottky diodes
                    · Highly doped regions : mesh dimensions
- Computation time remains long.

C.H.S.

LOW NOISE DEVICES

CHARGE CONTROL AND ANALYTICAL MODELS

- **Principle:**
  - One dimensional solution of equilibrium equations (Schrödinger or current continuity equations) following the transversal direction
  - Analytical formulations of drain current such as very simplified one: $J_D = q n_s <v>$

- **Interest:**
  - Usable for first order optimization
  - Allows to point out interesting physical dependances
  - Needs very small computation times.

- **Problems:**
  - Validity of simplified assumptions : unrealistic values of saturation velocity
  - Unable to treat complicated structures.

APPLICATIONS OF SIMPLIFIED CHARGE CONTROL MODEL: COMPARISONS WITH QUASI 2D MODEL
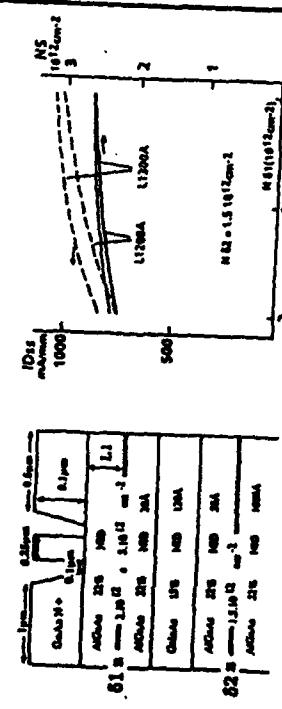
Considered structure

Charge control model
Noise software

From Cressler et al, 1992
ESPRIT 6835 AIMS

iemn

NOISE CALCULATIONS FOR F.E.T.

Noise origins :

- Carriers velocity fluctuations in the channel that induce
  - drain current fluctuations $<i_d>^2$
  - gate current fluctuations $<i_g>^2$

  that are correlated

- Noise in the access resistances Rs and Rg.

Circuit transformations for the noise determination

C.H.S.

GaAs MESFET : Basic principe.

Electrical equivalent circuit.

Slide 1:

CALCULATIONS OF GATE AND DRAIN NOISE SOURCES

1) In a quasi 2D scheme, representation of a noisy slice by a noisy source in proportionnal to $n(x)\, D(x)\, \Delta f$

(Diagram: Gate — noisy slice, x, x+Δx; Gate — noiseless slice, x, x+Δx, i_n)
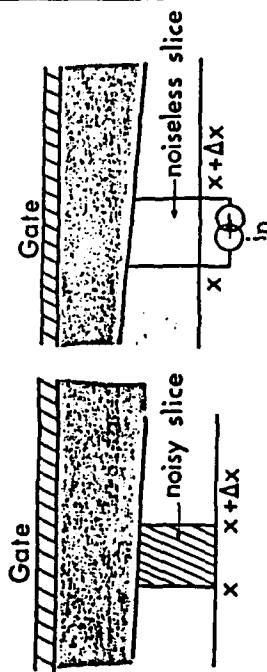
2) Use of the Impedance Field Method
(Van der Ziel, Schockley, Nougier, Cappy, Wang, Ghione)

- Calculation of transfer small signal impedances $Z(\omega, x)$ between noise sources in the channel and drain and gate voltages.

- Summation over the whole length of the channel.

Remarks:

- Interest of taking into account the distributed character of the channel (Cappy et al, 1990)

- Possibility of using a complete two dimensional treatment (Ghione 1990).

Slide 2:

ANALYTICAL FORMULATIONS OF NOISE FIGURE following Cappy

. When neglecting gate noise $<i_g>^2$:

● $F_{min} = 1 + 2 \sqrt{P}\, \dfrac{f}{f_c} \sqrt{g_m(R_s + R_g)}$ close to "Fukui" formula

with $P = \dfrac{<id^2>}{4kT g_m \Delta f}$

● According to Delagebeaudeuf : $P \# \dfrac{I_{d0}}{E_c L_g g_m}$

$F_{min} = 1 + 2 \sqrt{\dfrac{I_{d0}}{E_c L_g}}\; \times \dfrac{f}{f_c} \sqrt{R_s + R_g}$

● Consequences : For improving noise figure

- Rs and Rg must be reduced

- $f_c \# \dfrac{<v>}{2\Pi L_g}$ must be as high as possible.

● Remarks :

1) The intrinsic noise figure Fint (for $R_s + R_g = 0$) remains always close to 1 !!!

2) For comparing devices of similar technologies (L_g, R_s, R_g), the quality factor can be the ratio : $r = \dfrac{I_{d0}}{f_c^2}$

- When taking into account gate noise, more complicated formulae with Fint given by

$$F_{int} = 1 + 2\, \dfrac{f}{f_c} \sqrt{PR(1 - C^2)}$$

NOISE CALCULATIONS : PROBLEMS

iemn

- **Physical basis** : uncorrelation between the noise sources in the channel? Does it remain valid for very short gate length devices ($L_g \cong 500$ Å)?

- Values of diffusion coefficient are not well known : for instance in AlGaAs/GaAs heterojunction.

Reduced Diffusion coefficient vs Field in kV/cm

Legend: Fauquem. (1980), Bareikis (1980), de Murcia (1990), Gasquet (1985), Cast (1991)
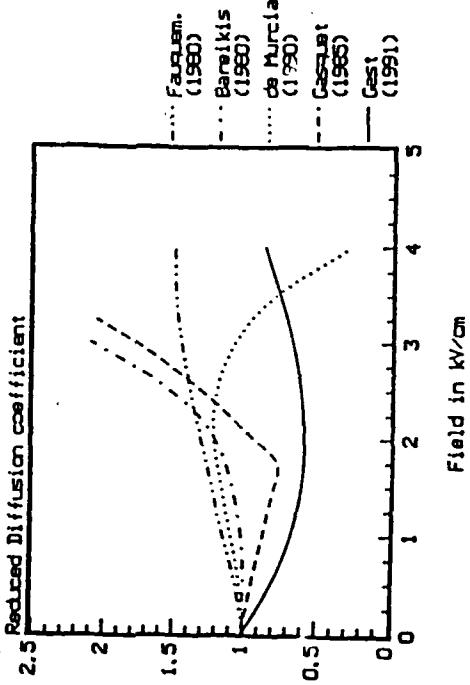
From Zimmerman et al, 1991

C.H.S.

Noise Performance of High Electron Mobility Transistors (HEMT)

SOURCE   DRAIN

n+ GaAs

SI-doped $Al_x\ Ga_{1-x}\ As$ x ~ 0.2

undoped $Al_x\ Ga_{1-x}\ As$ x ~ 0.2

electron channel undoped $In_y\ Ga_{1-y}\ As$ y ~ 0.2

GaAs

SI - GaAs

Pseudomorphic AlGaAs-InGaAs-GaAs HEMTs

Lg = 0.1 µm

* experimental data (Kao 1989 and Tan 1990)

— theory (Cappy 1990)

Minimum Noise Figure (dB) vs F (GHz)

C.H.S.

GaAs MESFET INFLUENCE OF THE ASPECT RATIO

$L_g = 1\mu$

$L_g = 0.4\mu$

$L_g = 0,1\mu$
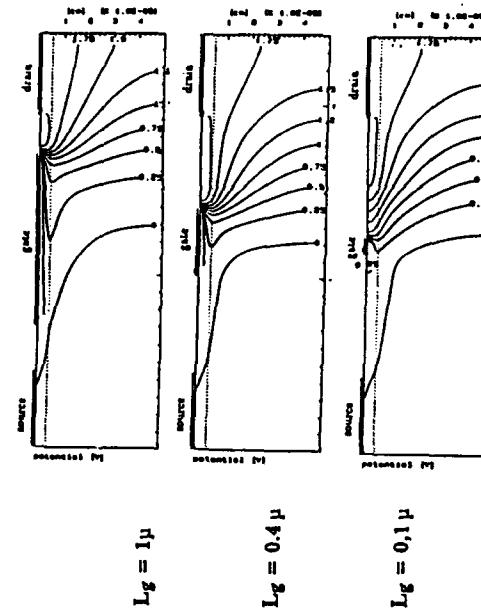
Equipotential contours

L_g μm	L_g/a	g_m mS/mm	g_d mS/mm
1	19	320	25
0.4	7.6	420	45
0.1	1.9	380	60

Simulations performed by Makowikz and Brockerhoff.
(Europ. Trans. on Tel.N°4 19 90).

LOW NOISE MESFET's

- Improvement of noise figure

  - Intrinsic $f_c = \dfrac{g_m}{2\Pi C_{gc}} \# \dfrac{<v>}{2\Pi L_c}$ must be increased

  - Main possibility. A decrease of gate length L_g that allows also an increase of $<v>$

Drift velocity 10^7 cm/s

.25µm

.5µm

1µm

Source   Gate   Drain

Typical variations of drift velocities in MESFETs for various gate lengths (Carnez et al)

But if the aspect ratio L_g/a is not high enough, short channel effects may cause a degradation of performances :

When L_g decreases g_m may decrease } Gain ↗
                        g_d increases
                        f_c is not improved

C.H.S.

RECESSED GATE GaAs MESFET

Gm
mS/mm
200
100

.1 .2 .3 .4 .5 .6 .7 R µm

Idss/2

.15Idss

Gd
mS/mm
20
10

.1 .2 .3 .4 .5 .6 .7 R µm
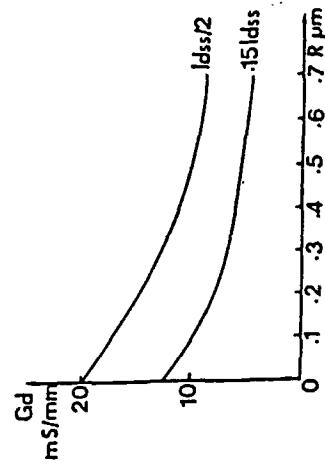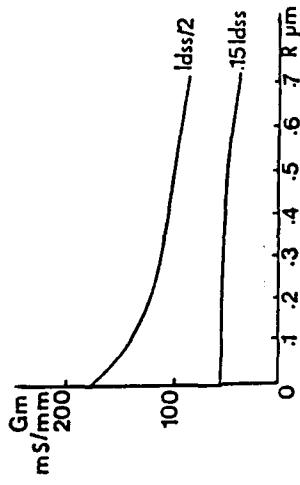
Idss/2

.15Idss

Transconductance and output conductance dependances upon the distance R for a 0,3 µm gate MESFET at $I_{dss}/2$ and 0,15 I_{dss}

Heliodore et al IEEE Trans. 1988.

SUPPRESSION OF SHORT CHANNEL EFFECTS

• Use of a partially depleted p layer:

NiGe/Au   WSi   NiGe/Au

n+   p   n+

depleted region

Schematic cross sectionnal view of the MESFET.

1000
800
600
400

200

100
g_m (mS/mm)

.5   1   2   5

S^+ : 50 KeV, 3.7E12 cm^{-2}
Be^+ : 70 KeV, 2E12 cm^{-2}

Vgs = 0.65V

Gate length dependance of g_m.

From Noda et al (IEEE Trans. 1991).

C.H.S.

Theoretical velocity field characteristics at 77° K

CONVENTIONAL AlGaAs/GaAs MODFET's
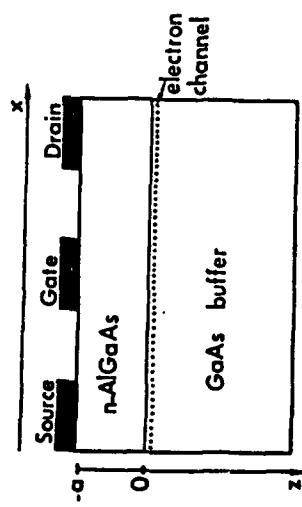
- Device Structure

- Band diagramm at equilibrium

CHS Lille

MESFET's CAPABILITIES :
Recent results
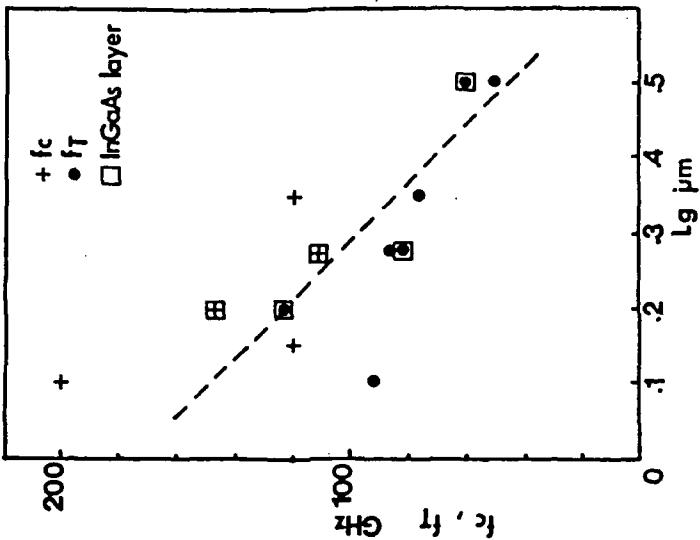
Current gain cut-off frequency versus gate length.

- 1991 -

VERY SHORT GATE LENGTHS MESFET's

Other possibilities for improving the performance :

- Unsymmetrical recessed gate structures $\frac{g_m}{g_d}$ can be improved by a factor of two

- Use of highly doped (ion implanted) InGaAs unmatched layers (Feng et al)

- Use of very highly doped (4.10^{18}) thin GaAs layers (150 Å) in order to maintain a good aspect ratio even for small gate length.

but carriers mobility decreases strongly down to 1 000 cm²/V sec.

PHYSICAL BEHAVIOUR OF CONVENTIONAL MODFET's
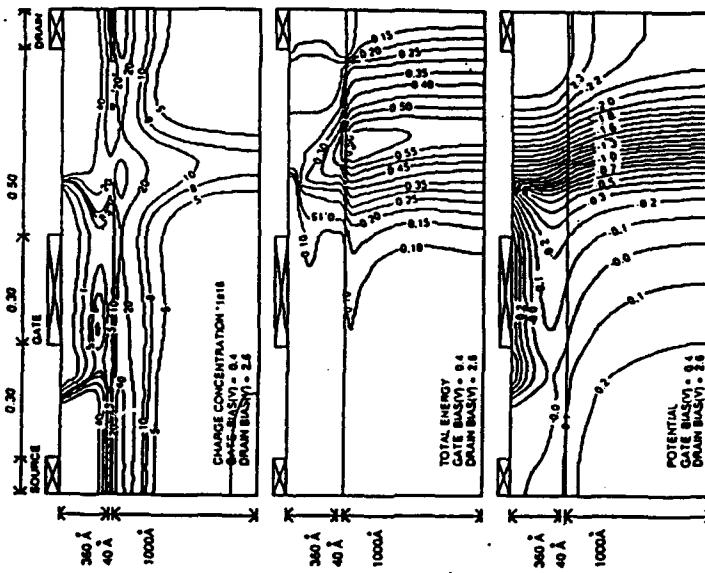
- Equiconcentration. Equienergy and equipotential contours for a 0.3 μm gate MODFET by taking into account surface effects

(from Shawki and Salmer)

CHS Lille

SUBBAND STRUCTURE IN THE QUANTUM WELL OF A CONVENTIONAL MODFET

- $E_k = E_{nx} + \dfrac{\hbar^2}{2m^*}(kx^2 + ky^2)$
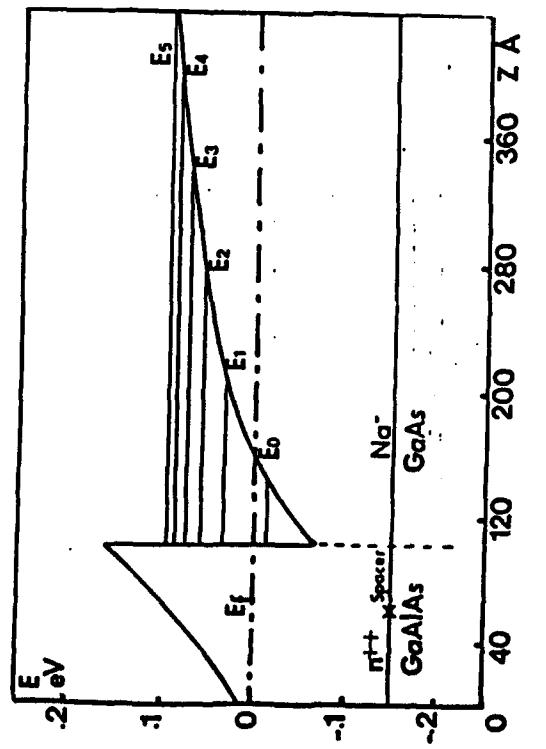
- If space charge reaction is neglected, and well triangular shape is assumed :

$$E_{nx} = \left(\dfrac{\hbar}{2m}\right)^{1/2}\left(\dfrac{3}{2}\,\Pi q\,E_s\right)^{2/3}\left(n + \dfrac{3}{4}\right)^{2/3}$$

Note that E_s is proportionnal to n_s.

CHS Lille

USE OF AN AlGaAs BUFFER LAYER

CURRENT LINES (mA/mm)

Conventional MODFET
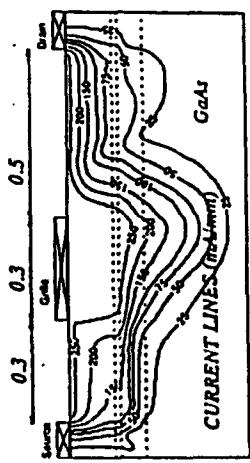
CURRENT LINES (mA/mm)
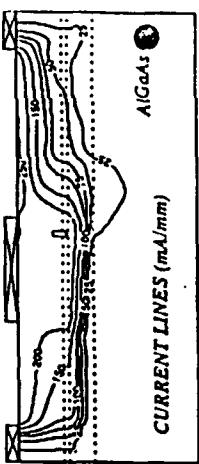
AlGaAs

AlGAAS Buffer Layer

Current lines in submicron gate MODFET's

Following Shawki and Salmer (1990)

PARTICULARITIES OF CHARGE CONTROL LAW IN CONVENTIONAL MODFET's

Variations of total capacitance C_g as function of V_g

Contributions of electrons

C_0 in the well
C_1 free in ALGaAs
C_2 trapped in AlGaAs

CHS Lille

CONVENTIONAL MODFET'S
Main limitations

- Total carrier concentration in the channel :

  $n_s < 1.2 \ 10^{12}/cm^2$

  $I_{dss} < 250 \ mA/mm$

- Quantum well height : $\Delta Ec < 0,25 \ eV$

- Carrier injection into the buffer layer

  → High values of gd.

- Threshold voltage variations.

- Parasitic effects under low temperature conditions.

---

iemn

SHORT CHANNEL EFFECTS IN CONVENTIONNAL HEMTS

Following Han and Ferry (1990), it is possible to define an effective gate length

$$L_{eff} = L_g + 2(a + \Delta d)$$

a   AlGaAs layer thickness
Δd equivalent distance of the 2D EG.

$50 \ Å < \Delta d < 100 \ Å$
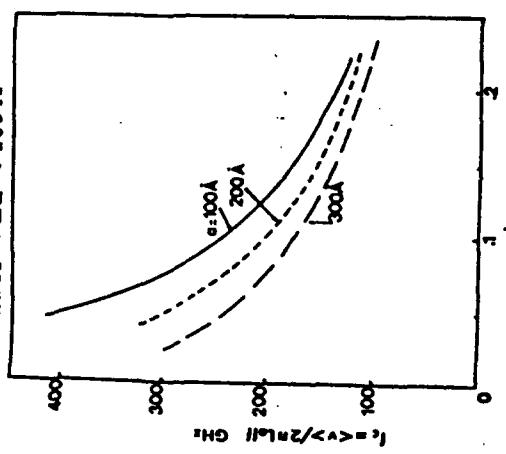
First order evaluation of the influence of aspect ratio on intrinsic current gain frequency.

IMPORTANCE OF THE ASPECT
RATIO: L_g/a

1/Lg law

$1/L_g$ μm^{-1}

f_c GHz
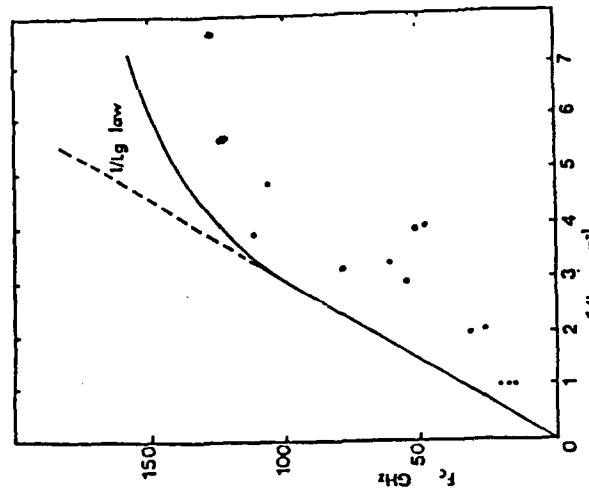
Theoretical intrinsic current gain cut-off frequencies ($f_c = g_m/2\Pi C_{gs}$) dependence upon gate length L_g, for a conventional MODFET, the AlGaAs layer thickness a remaining constant (a = 400 Å)

Following Fauquembergue calculations.

• Experimental extrinsic current gain cut-off frequencies.

CHS Lille

n⁺ GaAs   n⁺GaAlAs   2DEG   Contact
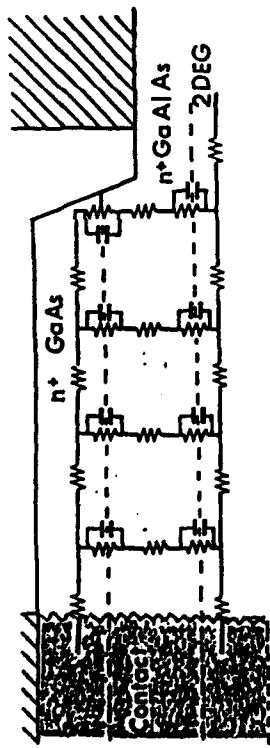
Equivalent scheme of the source access zone.

R
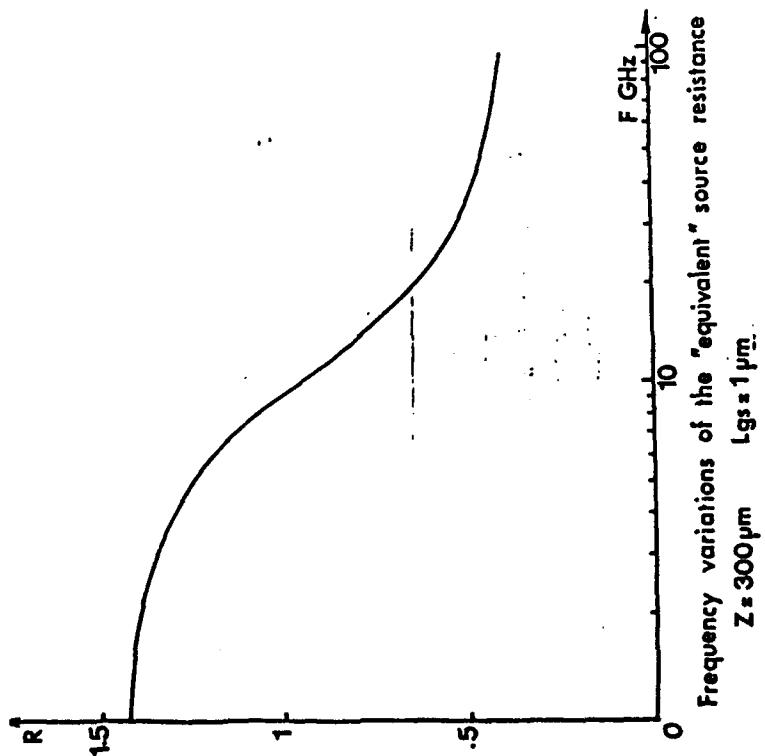
Frequency variations of the "equivalent" source resistance
Z = 300μm   Lgs = 1μm

F GHz

PLANAR DOPED MODFET :
Improvement of the transconductance
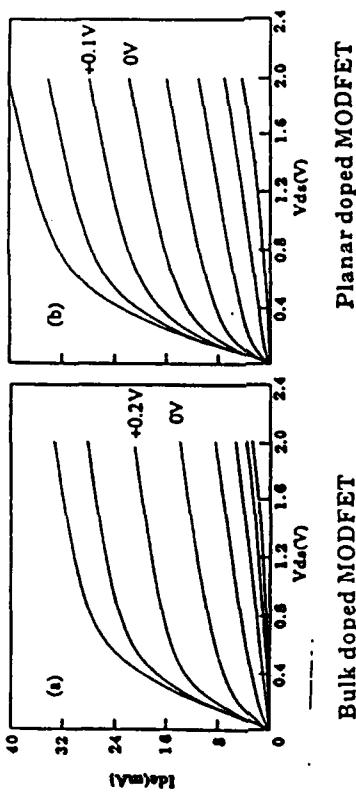
Bulk doped MODFET   Planar doped MODFET

Jin et al (1991)
C.H.S.

PLANAR DOPED HETEROSTRUCTURES

- Higher carrier concentration : $n_s \simeq 1,5\ 10^{12}$
- Improvement of the ratio L_g/a and then g_m/g_d (gain).
- Improvement of the breakdown voltage.
- Suppression of parasitic effects due to DX Center.
- But : Parasitic access resistances may increase.

PSEUDOMORPHIC AlGaAs/InGaAs MODFET's

Al Gate

n-Al.15 Ga.85 As

Undoped

In$_y$Ga$_{1-y}$As Undoped

p$^-$ GaAs

E$_c$

E$_f$
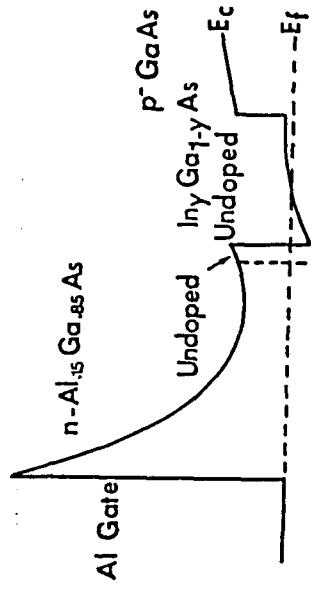
- Energy band diagram

Critical thickness Å — 800, 600, 400, 200

In Mole fraction y — 0 .1 .2 .3 .4
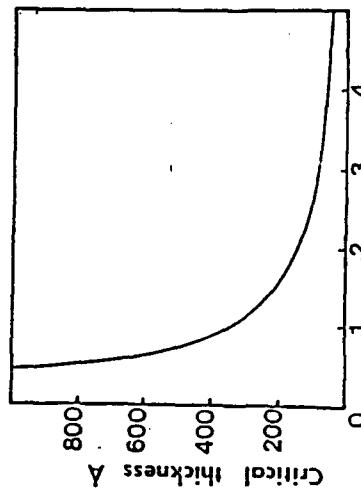
- Critical thickness of the well as function of In mole fraction

---

μm — .6 1 2 5 10

AlSb

GaSb

InAs

AlInAs

InP

Ga$_{.35}$In$_{.65}$As

CdS

ZnSe

GaAs

AlAs

In$_{.15}$Ga$_{.85}$As

Ga$_{.47}$In$_{.53}$As

Ge

a$_0$ Å — 5.6 5.8 6 6.2

E$_g$ eV — 3 2 1.45 .75
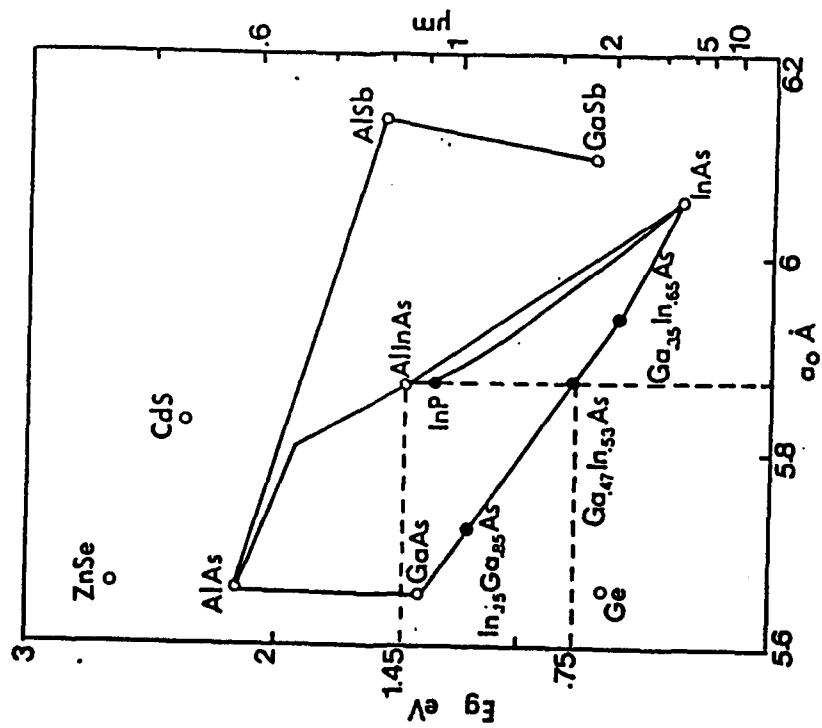
Energy band gap versus lattice constant.

PSEUDOMORPHIC AlGaAs/InGaAs MODFET's

Ns x E12/CM²

2,5
2,0
1,5
1,0
0,5
0,0

% In

0,0   0,1   0,2   0,3   0,4   0,5

% Al
,25
,2
,15
,1
,05
0

Carrier concentration dependence upon Al and In contents :
Self consistent solution of Schrödinger equation

From Cappy et al (1989)

PSEUDOMORPHIC AlGaAs/InGaAs
MODFET'S

Source   Gate   Drain

Si

n⁺GaAs
n⁺ AlGaAs
AlGaAs spacer
InGaAs channel
GaAs buffer

Advantages :

- More favourable transport properties ?

- Better confinement.

- Higher carrier concentration values :

ns = 2.5 10¹²/cm2.

CHS Lille

COMPARATIVE VARIATIONS OF INTRINSIC CAPACITANCE C_{GS}

Cgs pF/mm

Vgs v

Pseudomorphic AlGaAs/InGaAs/GaAs ———

Conventional AlGaAs/GaAs - - - -

$L_g = 0,3 \ \mu m$ $N_D = 10^{18}$ $x_{Al} = 0,2$

$y_{In} = 0,2$

CHS Lille

PSEUDOMORPHIC MATERIALS
Influence of strain

x=0.25 unstrained
GaAs unstrained
x=0.25 strained

Electric field (kV/cm)

Velocity (10^7 cm/s)

Velocity field characteristics
in InxGa1-xAs layers

From Thobel et al (1989)

PSEUDOMORPHIC AlGaAs/InGaAs MODFET'S

• Various structures :

- n channel - uniform doping
- n channel - planar doped
- n multiple channel
- n short periode superlattice channel
- n channel with variable In content
- inverted device
- MIS Like FET
- p channel

• Among the best performance :

- Tan et al (TRW)

L_g = 0,1 µm f_T = 135 GHz
N_F = 2.1 dB G_a = 6.3 dB at 94 GHz.

C.H.S.

PSEUDOMORPHIC INVERTED MODFET
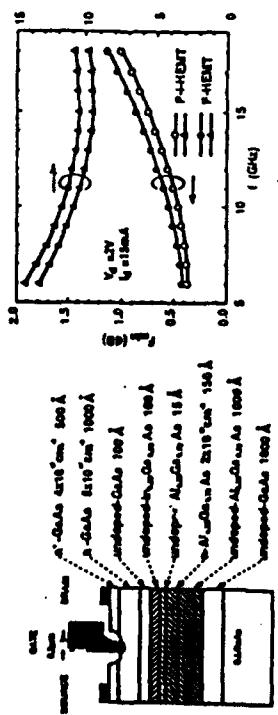
Schematic cross-section of the device
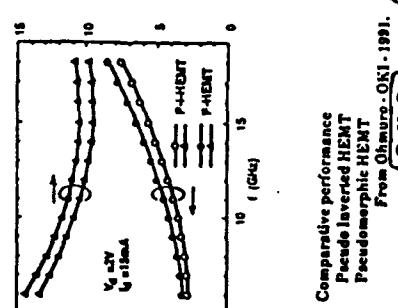
Comparative performance
Pseudo inverted HEMT
Pseudomorphic HEMT

From Ohmuro - OKI - 1991.

C.H.S.

LOW NOISE PSEUDOMORPHIC MODFET's
EXEMPLE OF REPRODUCTIBILITY AND FIABILITY
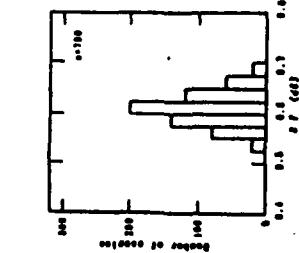
Distribution of noise figure at 12 GHz
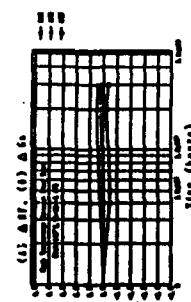
NF as function of test time.

From Tokue - IEDM 91

C.H.S.

INFLUENCE OF QUANTUM EFFECTS
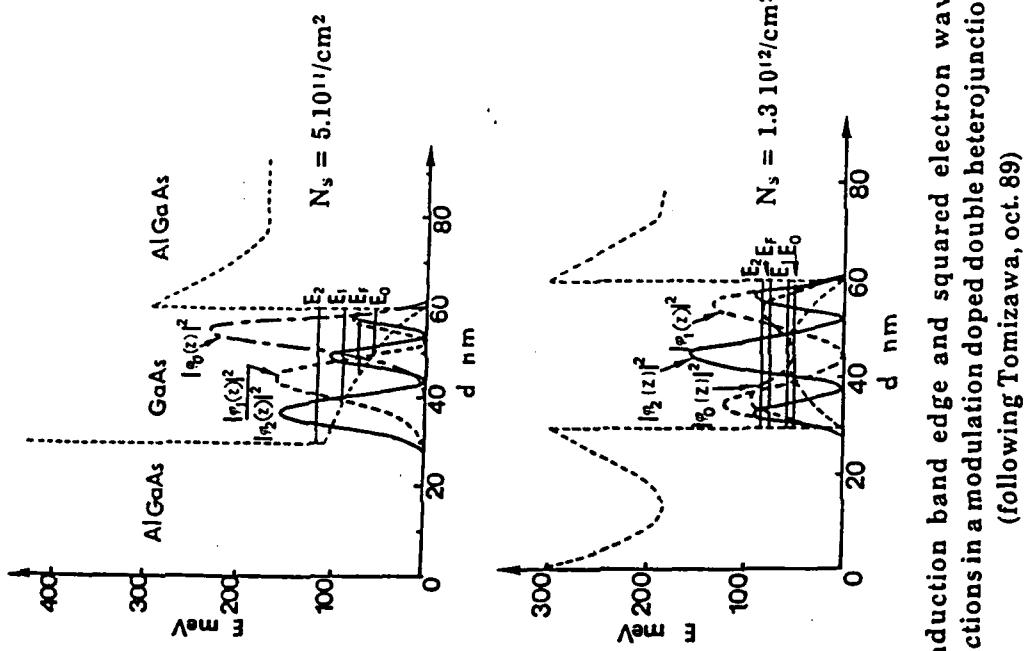
Conduction band edge and squared electron wave functions in a modulation doped double heterojunction (following Tomizawa, oct. 89)

CARRIER MOBILITY VARIATIONS IN AlGaAs/InGaAs PSEUDOMORPHIC STRUCTURES

SHEET CARRIER DENSITY (1E12/cm2)

Gated Van der Pauw Hall Effect Measurements

From Pons - Thomson CSF.

C.H.S.

ADVANTAGES OF InAlAs/InGaAs on InP MODFETS

μ 10⁴cm²/V.s

3
2.5
2
1.5
1
.5
0

.2 .4 .6 .8 1
InAs mole fraction

---- Strained on InP
—— Unstrained
·-·- Strained on GaAs

f_c GHz
150
100
50
0

2 on InP
1 Conventional

-2 -1.5 -1 -.5 0
$V_{gs}-V_{bi}$ v
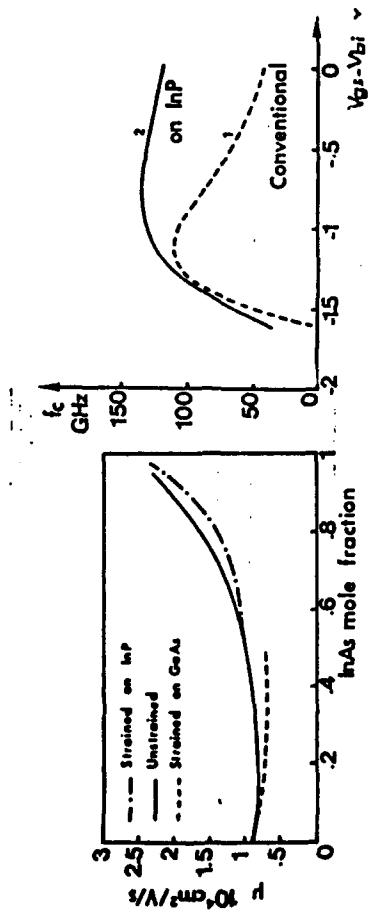
Improvements of carrier mobility and cut-off frequency

MODFET on InP SUBSTRATE

Source — TiPtAu Gate

Drain — AuGeNiAgAu Ohmic Contact

Si-Planar Doping

InAlAs Spacer

n⁺ InGaAs

InAlAs

InGaAs

InAlAs

InP Substrate
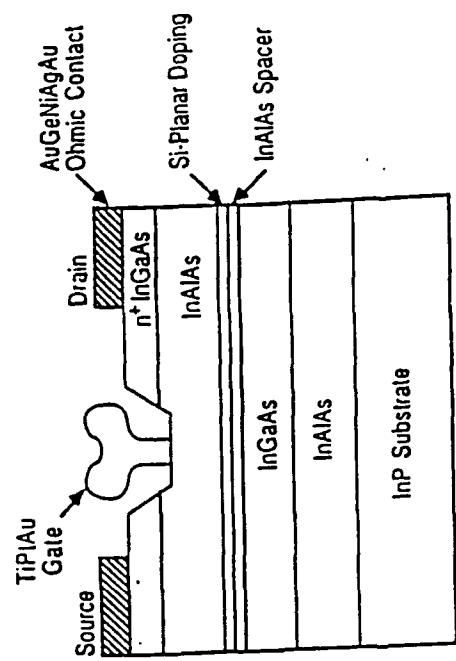
Schematic cross section of a
$In_{0.52}Al_{0.48}As/In_{0.53}Ga_{0.47}As$ MODFET

Potential interest

. Better velocity field characteristics
. Higher conduction based discontinuity
  and then better confinement
. Higher carrier concentration in the
  channel.

INTEREST OF InGaAs ON InP PSEUDOMORPHIC STRUCTURES

- Conduction Band Structure

- Ga$_{1-x}$In$_x$As/Al$_{.48}$In$_{.52}$As Band offsets.

- Electron effective mass in Ga$_{1-x}$In$_x$As strained on inP substrate
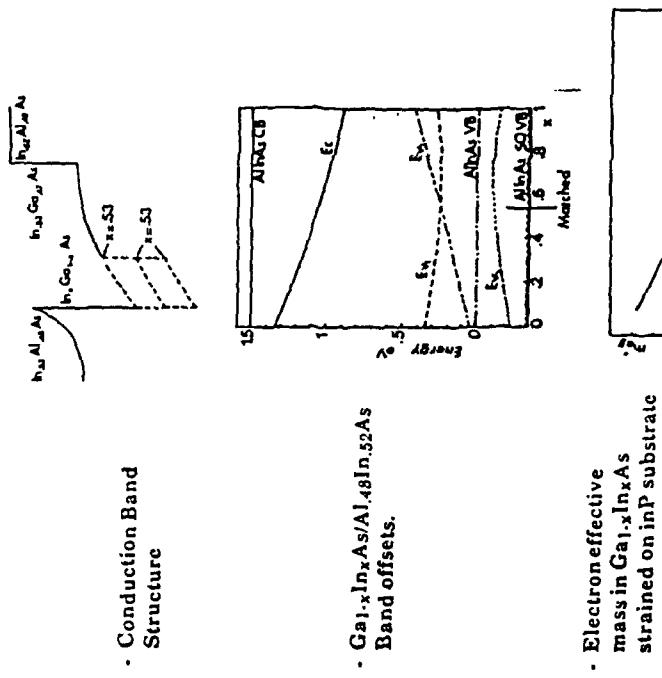
Following Priester (1990).

C.H.S.

MULTIVALLEY TRANSFER ACROSS THE HETEROJUNCTION

$\Delta E_{LL} = 1.13\,eV$

$\Delta E_{XX} = .43\,eV$

$\Delta E_C = .52\,eV$

In$_{.53}$Ga$_{.47}$As

In$_{.52}$Al$_{.48}$As

L   X   G

iemn

INFLUENCE OF CAP LAYER CHARACTERISTICS
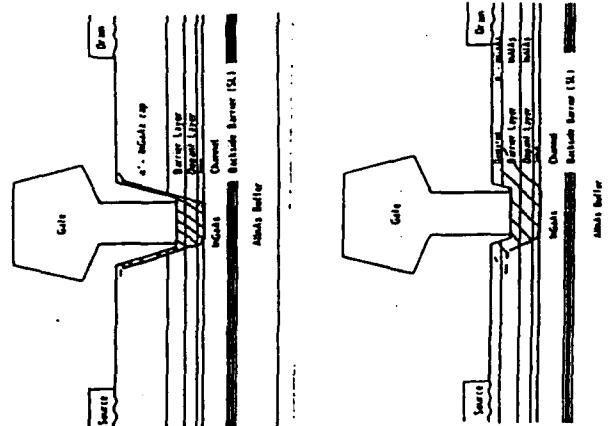
Trade-off between :

- Improvement of parasitic resistance :
  Thick-Doped.
- Improvement of output conductance :
  Thin and depleted cap layer

From Dickman (1991).

C.H.S.

---

iemn

$In_{.52}Al_{.48}As/In_xGa_{1-x}As$ on InP PSEUDOMORPHIC STRUCTURES

on I.M. GaInAs

Tg=450°C
Tg=525°C

on rough InP
Tg=525°C

In composition

Critical thickness A

Critical layer thickness   as function of I_n content

Victorovitch et al (1991)

293K

InAlAs/InGaAs/InP

In content

Electron mobility m²/Vs

Lattice match

Electron mobility   as function of I_n content

Tacano et al (1991)

C.H.S.

HIGH PERFORMANCE InGaAs on InP MODFET

iemn

- 0,1 μm T gate Pseudomorphic (In content = 60 %)

Tan et al IEDM 91

200 Å 4E10 cm⁻³	InGaAs(x=.60)
200 Å undoped	InAlAs(x=.52)
250 Å undoped	InGaAs(x=.60)
AlInAs Buffer(x=0.52)	
Superlattice Buffer	
S.I. InP Substrate	

$f_T = 200$ GHz
$f_c = 260$ GHz

$N_F = 1,3$ dB
$G_a = 8.2$ dB   } at 95 GHz
$G_a = 7.3$ dB at 141 GHz

Device cross section

- 650 Å Selfaligned gate Pseudomorphic (x = 80 %)

Nguyen et al. IEEE El. Dev. March 92

Best current gain
cut-off frequency
$f_T = 305$ GHz

Transconductance values:
Lattice matched versus
pseudomorphic devices.

C.H.S.

IMPROVEMENT OF THE OUTPUT CONDUCTANCE AND DECREASE OF KINK EFFECT IN InGaAs/InAlAs MODFET (Zhan et al) 1991.

iemn

Main cause : Ionization in the channel

Gate — Lg — Drain

Schottky layer — n + InGaAs
Carrier supply layer — I InAlAs
Channel layer — n + InAlAs
Buffer layer — InGaAs
Substrate — I InAlAs
— S.I. InP
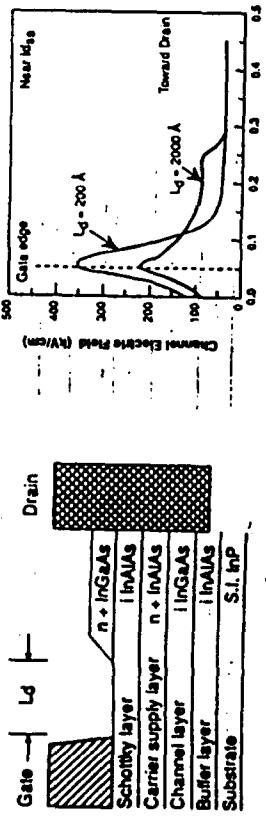
Schematic cross-section
of the device

Electric field in the
channel deduced from simulation

C.H.S.

FIELD EFFECT TRANSISTORS

State of the art (March 1992)

+ GaAs and InGaAs MESFET
◄ Strained InAlAs/InGaAs MODFET
○ Lattice Matched InGaAs on InP MODFET
✱ Pseudomorphic AlGaAs/InGaAs MODFET
■ AlGaAs/GaAs MODFET

MODFET's ON InP SUBSTRATE :
NOVEL STRUCTURES

- InAlAs/InAs : Eugster et al - Dec. 91
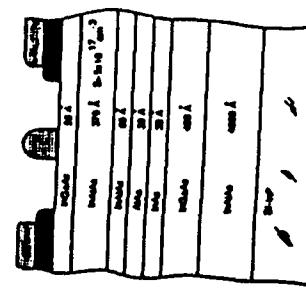
  <u>Remarks</u> : Very thin InAs layer for avoiding dislocations.

  Small mobility : 5400 cm2/Vsec

  Reasons are not clear :

  - Strongly quantizied channel ?
  - Dislocations ?
  - Roughness of interfaces.

  $L_g = 2\,\mu m$ $g_m = 170\ mS/mm$

- InAs/AlSb - Werking et al (1992) - Yo et al (1991)

  - Very promising properties

  $n_s = 3.8\,10^{12}$
  $\mu = 21\,000\ cm2/Vsec$

  For $L_g = 2\,\mu m$
  $g_m = 473\ mS/mm$

  - Very difficult technologies

Energy band diagramm

POWER FET

C.H.S.

LOW NOISE FIELD EFFECTS TRANSISTORS

Noise figure versus frequency
Among the best results
published (March 92).
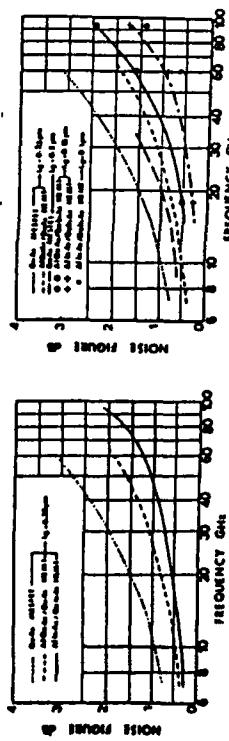
Comparison between FETs
for the same gate length (0.25 μm)
from DUH (MTT 89)

C.H.S.

COMPARATIVE BEHAVIOUR OF 0.2 μm GATE FET'S
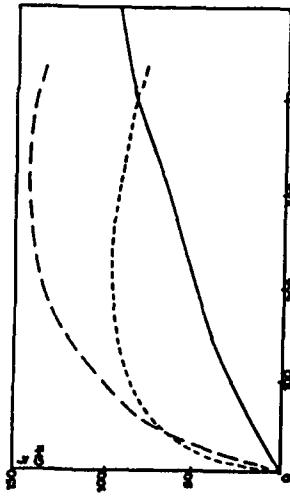
· GaAs MESFET · very well optimised (from PONS) – – – – – –
· not optimised ————
· Pseudomorphic AlGaAs/InGaAs/GaAs MODFET ————

C.H.S.

LARGE SIGNAL SIMULATIONS OF POWER FETs

It is necessary but very <u>difficult</u> to take into account at the same time :

- Physical device behaviour (hot electron effects, surface effects) and physical limitations (breakdown)
- Harmonic components of drain voltages and currents

forward biasing limit
I_{DSS}
Drain current, Id (mA)
100
50
0
Vgs Volts
-0.5
-1
fundamental
total
-1.5
-2.5
V_{BR}
voltage limit
0 5 10
Drain-Source voltage, Vds (Volts)

from Halkias (1988).

- Parasitic elements
- Device configuration : association of a large number of gates (100 at 100 GHz ?)
- Thermal considerations (dc and pulse).

POWER FET LIMITATIONS

P_{IN} V_{GS} P_{OUT} I_D V_{DS} Load C_{OUT} R_{OUT} Load

I_D mA
I_{DSS}
100
$V_{GS}+0.5v$ -1v -2v -3v
V_{BR} V_{DS} v
P_{OUT} W
10
P_{IN} W

- $P_{out} \# 1/8 \, I_{DSS} \, V_{BR} \cos \Phi$

<u>Requirements :</u>

- I_{DSS} } as high as possible.
- V_{BR}
- Output impedance less reactive as possible
- Gain linearity : variation of MAG with V_{GS}.

SIMULATIONS OF BREAKDOWN PHENOMENA

- Simplified models

- Quasi 2D models : (Wroblewski 1983)
  . Channel breakdown : good estimation
    Optimization of MESFET (recess structure)
  . Problems of gate breakdown, calculation.

- 2D Solution of Poisson equation (Wroblewki - Zaitlin, Fellon).

  . Calculation of multiplication factor.
  . Evaluation of electric field at the gate edges
    → Tunneling currents.
  . Problems :

  • knowledge of ionization rates and accurate predictions of tunneling currents.
  • space charge reaction is not included : valid only for small Ids.

- 2D models (Mizuta, Wada, Lefebvre, Chan, Barton, Ashworth, ...)

  . Introduction of ionization effects or tunneling currents.
  . Problems :

  • drift diffusion models
  • Ionization effects and tunneling currents are not introduced together.

C.H.S.

BREAKDOWN PHENOMENA : PHYSICAL KNOWLEDGE

- Example of MESFET
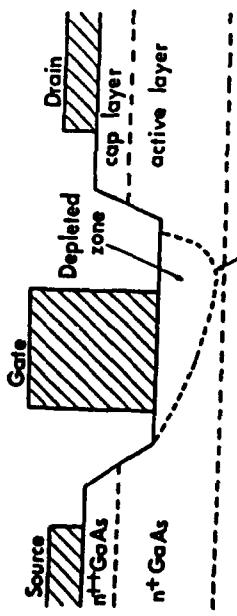
- Physical causes

  • Impact ionization in the channel:
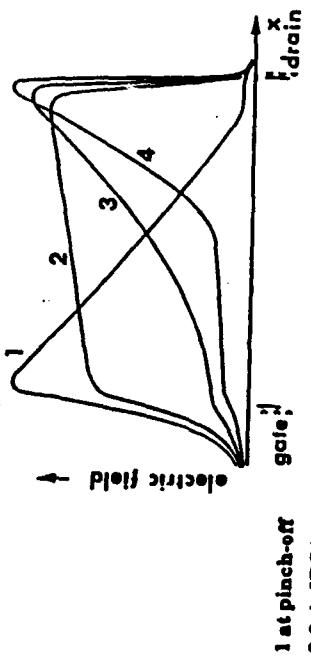
1 at pinch-off

2-3-4 : IDS increasing   From Wroblewski et al (1983)

  • In the gate depletion region
    - impact ionization (at pinch-off)
    - gate current (thermoionic or tunneling)

C.H.S.

iemn

BREAKDOWN VOLTAGE : MAIN DEPENDANCES

- **Material properties :**

  - Ionization rates in bulk material
  - Problems : Ionization rates in
    - Very thin channels ??
    - Strained materials : Improvement of the breakdown voltage due to strained : m^*h ↘ α

      (Eisenbeiser - 1991)

- **Epilayer characteristics :**

  - Doping level of the last layer under the gate is the most important.
    - Trade off with high Idss
    - Use of Low-High doping profile
  - Very thin top layer (100 to 200 Å). Influence of tunneling currents.

- **Semi insulated substrate :** Hole capture.

- **Recess zone configuration :**

  - Double recess gate process allow to improve breakdown voltage :
    from 8 to 16 V in InAlAs/InGaAs/InP HEMT (Boss - 1991).

  - Problems of surface states under dynamic conditions (Huang - 1991).

C.H.S.

---

iemn

IONIZATION RATES IN BINARY AND TERNARY III-V MATERIALS
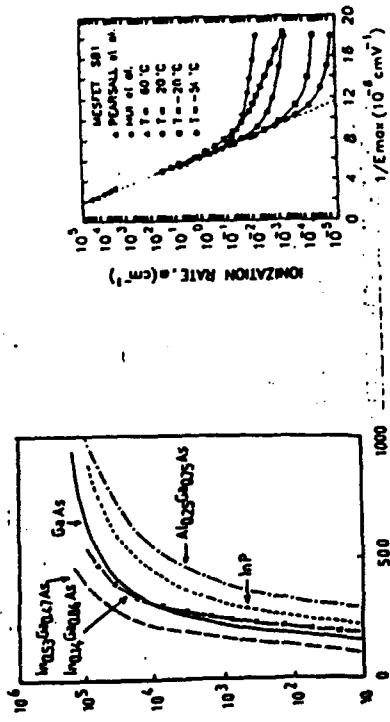
Ionization coefficients data for InGaAs

Accurate determination of α in GaAs by Canali et al (1991)

C.H.S.

=iemn

CALCULATIONS OF GATE CURRENT

- Main origins :
  - Thermoionic emission process (T.E.)
  - Field effect tunneling (F.E.)
  - Thermoionic field effect tunneling (T.F.E.)
  - Surface effects.

- Example of comparative values
  Current density versus electric field for a
  Schottky (following Barton).

BREAKDOWN IN HEMT : TUNNEL CURRENTS

CONVENTIONAL HEMTS

PULSE-DOPED HEMTS

Following CROSNIER et al

iemn

IMPROVEMENTS OF BREAKDOWN VOLTAGE AND PERFORMANCE BY USING A LOW-HIGH DOPING PROFILE (Takahashi - 1991).

$I_{MAX} = f$ (Breakdown V Structure: II $T_i = 0$.

Electric field distribution Structure I : $T_i = 300A$

Device Structure.

C.H.S.

---

iemn

IMPROVEMENTS OF GaAs MESFET

- **Pulse doped active layer :**

  . It allows to improve breakdown voltage and linearity : for instance Chu obtained 23 dB between IP3 and P0-1dB

  . Such improvement was predicted by Heliodore

  - Example of result : 220 mW output power at 35 GHz (RCA) 0.5 μm x 600 μm gate.

- **Highly doped InGaAs channels (Feng and Wang)**

  $L_g = 0.25$ μm

  96 mW at 44 GHz   0.6 W/mm in

  121 mW at 60 GHz   both cases

- **Pulse doped InGaAs channel : Kim**

  $L_g = 0.25$ μm $I_{DSS} = 700$ mA/mm

  $P_0 = 50$ mW - 0.6 W/mm at 60 GHz.

- **Other Improvements : AlGaAs buffer - via holes.**

C.H.S.

iemn

POWER FIELD EFFECT TRANSISTORS

● **MESFET :**

- GaAs
- InGaAs on GaAs substrate.

● **MODFET :**

- Conventional AlGaAs/GaAs
- Multichannel AlGaAs/GaAs
- Single channel AlGaAs/InGaAs
- Multichannel AlGaAs/InGaAs/GaAs
- InAlAs/InGaAs on InP substrate..

● **HFET or D.M.T. :**

- AlGaAs/InGaAs (n+ doped) strained or not on GaAs.
- InAlAs/InGaAs (n+ doped) on InP.
- InAlAs/InP (n+) on InP substrate.

● **MISFET :**

- InGaAs   n+ doped   } on InP
- InP   n+ doped
- LT GaAs insulating layer.

C.H.S.

SENSITIVITY OF THE MAXIMUM ELECTRIC FIELD AT THE GATE CORNER OF A PLANAR DOPED PM HFET TO THE STRUCTURE PARAMETERS.

extracted from the 2 D
electric field map under
pinch-off conditions
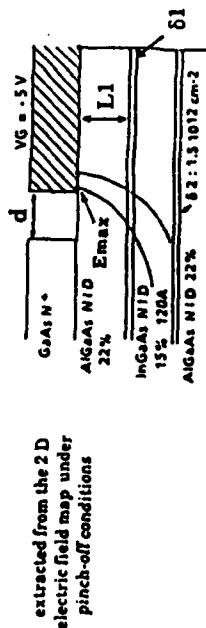
Influence of the
recess spacing d
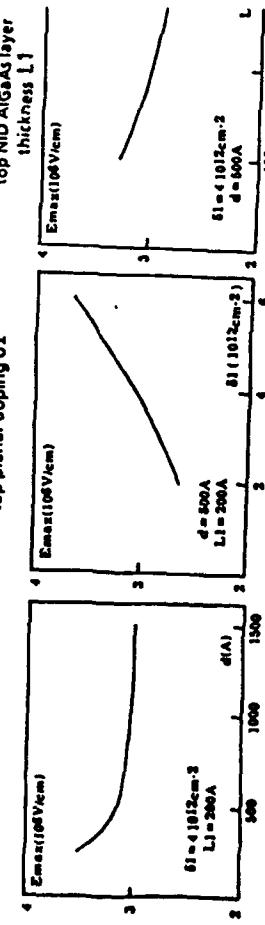
Influence of the
top planar doping $\delta 1$

Influence of the
top NID AlGaAs layer
thickness L1

IMPROVEMENTS OF GaAs MESFET

- Pulse doped active layer :

.It allows to improve breakdown voltage and linearity : for instance Chu obtained 23 dB between IP3 and P_{0-1} dB

.Such improvement was predicted by Heliodore

- Example of result : 220 mW output power at 35 GHz (RCA)
0.5 μm x 600 μm gate.

- Highly doped InGaAs channels (Feng and Wang)

$L_g = 0.25$ μm

96 mW at 44 GHz   0.6 W/mm in

121 mW at 60 GHz } both cases

- Pulse doped InGaAs channel : Kim

$L_g = 0.25$ μm $I_{DSS} = 700$ mA/mm

$P_0 = 50$ mW ⟶ 0.6 W/mm at 60 GHz.

- Other improvements : AlGaAs buffer - via holes.

C.H.S.

IMPROVEMENTS OF BREAKDOWN VOLTAGE AND PERFORMANCE BY USING A LOW-HIGH DOPING PROFILE (Takahashi - 1991).

| Device Structure. | Electric field distribution Structure I : $T_i = 300\text{Å}$ | $I_{MAX} = f$ (Breakdown V) Structure : II $T_i = 0$. |

C.H.S.

PSEUDOMORPHIC HFET LAYER STRUCTURES

HEMT

4 x 10^18 GaAs
1 x 10^17 Al.30Ga.70As
5 x 10^12 Si planar doping
undoped Al.30Ga.70As spacer
undoped In.22Ga.78As well 125 Å
Al.30Ga.70As/GaAs superlattice
GaAs buffer
Semi-insulating GaAs

$n_s = 2,5\ 10^{12}$
$\mu = 6\ 300$

DCHFET

4 x 10^18 GaAs
1 x 10^17 Al.25Ga.75As
2.6 x 10^10 In.22Ga.78As well 125 Å
Al.25 Ga.75As/GaAs superlattice
GaAs buffer
Semi-insulating GaAs

$n_s = 4,9\ 10^{12}$
$\mu = 1\,900$

DCH MODFET

4 x 10^18 GaAs
1 x 10^17 Al.25Ga.75As
5 x 10^12 Si planar doping
undoped Al.25Ga.75As spacer
undoped In.22Ga.78As well 63 Å
1,1 x 10^12 Si planar doping
undoped In.22Ga.78As well 63 Å
Al.25 Ga.75As/GaAs superlattice
GaAs buffer
Semi-insulating GaAs

$n_s = 4\ 10^{12}$
$\mu = 3\,100$

DH HEMT

4 x 10^18 GaAs
1 x 10^17 Al.25Ga.75As
4 x 10^12 Si planar doping
undoped Al.25Ga.75As spacer
undoped In.22Ga.78As well 125 Å
undoped Al.25Ga.75As spacer
1 x 10^18 Al.25Ga.75As 50 Å
Al.25 Ga.75As/GaAs superlattice
GaAs buffer
Semi-insulating GaAs

$n_s = 3,4\ 10^{12}$
$\mu = 5\,200$

From Lester et al (1989)

---

iemn

CONVENTIONAL AlGaAs/GaAs HEMT

- Single channel

- Sheet carrier density smaller than 1.2 10^12 cm2 : influence of DX centers for x ≥ 0.2

- IDSS is too small : < 300 mA/mm.

- Multichannel devices

  - ns and IDSS may be higher
    ns = 5.3 10^12/cm2
    IDSS = 1.100 mA/mm for a 3 channels.

- Improvement of linearity

- Problem of Rs
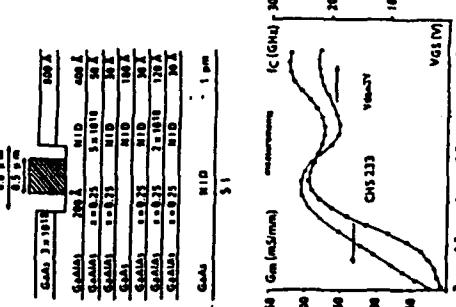
From Temcamani

- Example of performance : Fujitsu
  0.5 x 2 400 μm IDSS = 530 mA/mm
  P0 = 1.2 W at 30 GHz.

C.H.S.

iemn

AlGaAs/InGaAs PSEUDOMORPHIC HFET

DOPED CHANNEL OR NOT?

- Best result at 60 GHz : 1 W/mm - L_g = 0,25 µm
  Homogeneously doped channel f_M = 150 GHz.

- Work of Dickmann et al (1991)

  * Comparison
    - Conventionnal HEMT
    - Centrally pulse doped channel HFET
    - Backside pulse doped channel HFET

  * Experimental results : $f_M = f(I_{DS})$.

C.H.S.

PSEUDOMORPHIC HFET STRUCTURES

Summary of D.C. characteristics and 35 GHz Power Performance

DEVICE	HEMT	DCHFET	DHEMT	DCH MODFET
Current Density mA/mm	455	470	630	600
Transconductance mS/mm	625	745	790	747
Breakdown Voltage v	9.5	7.2	9.5	9.5
Efficiency %	40	35.5	43	47
Maximum Power W/mm	0.75	0.77	0.97	0.94

Following Lester et al - DRC 1989

HEMT ON InP SUBSTRATE FOR POWER APPLICATIONS

iemn

- AlInAs/InGaAs/InAlAs HEMT

- Typical device structure (Matloubian)

SOURCE   GATE   DRAIN

GaInAs — CAP
n+ AlInAs — SCHOTTKY
n+ AlInAs — DONOR
n+ GaInAs — SPACER
n- AlInAs — CHANNEL
AlInAs — SPACER
n- AlInAs — δ-DOPED
BUFFER

InP SEMI-INSULATING SUBSTRATE

Promising results

- Matloubian et al : L_g = 0.22 μm
  V_B = 9 V I_{DSS} = 770 mA/m
  P_0 = 0.96 W/mm at 12 GHz   η = 40 %.

- Kao et al : L_g = 0.12 μm. Pseudomorphic
  V_H = 4.8 V I_{DSS} = 600 mA/mm
  P_0 = 0.35 W/mm at 60 GHz
  P_0 = 0.2 W/mm at 94 GHz - 100 mW
  (as compared to 67 mW for InGaAs/GaAs HEMT)

- Kuang et al : D HEMT
  I_{DSS} = 1.5 A/mm.

But problems of reliability.

C.H.S.

HIGH POWER DENSITY GaAs MISFET WITH A L-T GROWN AS THE INSULATOR

Source   Gate   Drain

L.T. GaAs
GaAs N_D = 6 10^{17}
AlAs

1000Å
100Å

Buffer layer GaAs

Device Structure.

F = 1.1 GHz
Vgs = -2v
940mW
870mW

Power-added efficiency %

Vds = 18.5v
Vds = 17.5v
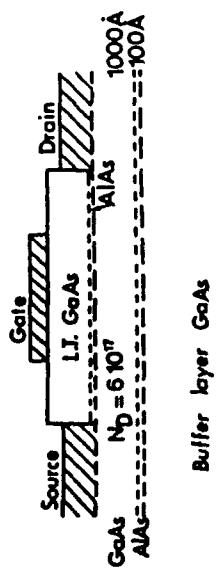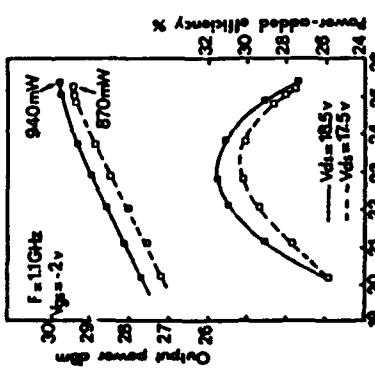
Power performance at 1,1 GHz
from Chen (1991).

iemn

EXAMPLE OF HIGH BREAKDOWN PSEUDOMORPHIC HEMT

Fi
Fmax

VDS (V)

Low Noise   Power

Current gain cut-off frequency and maximum gain cut-off frequency versus drain voltage : increase of transit time.

Schematic cross section of the gate drain region.

From Huang (1991).

C.H.S.

---

iemn

SCHRÖDINGER EQUATION SOLUTION → COMPARISON BETWEEN

Centrally pulse doped HFET and Backside pulse doped HFET

. Backside pulse doped is better
. But transfer to the E1 level occurs earlier.

From Dickman et al (1991).

C.H.S.

RAPPORT 1993 AVRIL

IEMN - CHS UNIVERSITE DE LILLE

1 D SCHRODINGER-POISSON SIMULATION OF THE ELECTRON GAS CONTROL NS = f(VG) IN PLANAR DOPED PSEUDOMORPHIC HFETS.

CONCLUSION

PHYSICAL MODELS:

- They are able to furnish very interesting informations on device behaviour;

- They can be used systematically for device optimization;

- They cannot give always accurate predictions due to the lack of knowledge in basic physics.

LOW NOISE DEVICES:

- Pseudomorphic AlGaAs/InGaAs MODFET can now constitute the basis of a mature technology;

- InGaAs on InP MODFET is the most promising but problems of reliability must be solved;

- Noise figure smaller than 1 dB at 94 GHz can be hoped in near future;

- Temperature dependance must be studied.

POWER DEVICES:

- The situation is not clear, but for next future, MODFETs on GaAs remain preferable;

- Studies must be done on several kinds of devices. New device configurations must be envisaged for adding properly output powers of devices.

C.H.S.

OTHER POWER FET DEVICES ON InP SUBSTRATE FOR MILLIMETER WAVE RANGE

HFET:

- InAlAs/n+ InGaAs   Del Alamo
  - Interesting results but trade-off between V_B (max 19 V) and IDSS (max 450 mH/mm)
  - No power results available.

- InAlAs/n+ InP   Del Alamo - Aina
  - Promising results in term of g_m and IDSS for long devices.
  - No power results.

MISFET:

- InP: - Very high breakdown voltage
  - High drain current density
  - Small gain
  - Problems of stability.

Very promising results:
  . 1.6 W/mm at 18 GHz. Bredenbender
  . 1.8 W/mm at 30 GHz. Saunier et al.

- InGaAs: - High drain current density
  - Small gain.

Encouraging results: Johnson - 1991.
L_g = 1 µm P_0 = 1.07 W/mm at 12 GHz

C.H.S.

TWO - TERMINAL MILLIMETER - WAVE DEVICES

J. Freyer

University of München • Germany

Two terminal millimetre-wave devices

J. Freyer, Technical University of Munich

- Basic principles of power generation
- Gunn element (Transferred electron device)
- IMPATT diode
- MITATT and TUNNETT diode
- PIN-avalanche diode
- BARITT diode
- Double-barrier and QWITT diode
- Device fabrication and packaging
- Resonators
- Application and state-of-the-art performance

$Z_D + Z_L = 0$ Kirchhoff's law

$R_D + R_L = 0$

$X_D + X_L = 0$

$P_{rf} = 1/2 \, |I_{rf}|^2 R_L = 1/2 \, |I_{rf}|^2 \, |R_D|$

(available output power)

series resistance R_s

$P_{rf} = 1/2 \, |I_{rf}|^2 \, (|R_D| - R_s)$

Basic equations

$$\frac{dn}{dt} = \frac{1}{q}\frac{dJ_n}{dx} + G$$

$$\frac{dp}{dt} = -\frac{1}{q}\frac{dJ_p}{dx} + G$$

continuity equations

$$J_n = qv_n n$$

$$J_p = qv_p p$$

current density equations

$$\varepsilon \frac{dE}{dx} = q(N_D - N_A + p - n)$$

Poisson equation

n : electron density
p : hole density
J_n : electron current density
J_p : hole current density
G : generation
q : electronic charge
t : time

x : space coordinate
v_n : electron drift velocity
v_p : hole drift velocity
ε : dielectric constant
E : electric field
N_D : donor density
N_A : acceptor density

Schematical presentation of a transit-time device

Voltage, injected convection current and induced current of a transit-time device

n-type semiconductor

$\dfrac{\partial E}{\partial x} = -\dfrac{q}{\varepsilon}(n - N_D)$ Poisson equation

$I_c = Aqnv(E)$ convection current

$I_{tot} = I_c + A\varepsilon\dfrac{\partial E}{\partial t}$ total current

$I_{tot} = A\left(\varepsilon\dfrac{\partial E}{\partial t} + q(N_D\dfrac{\varepsilon}{q}\dfrac{\partial E}{\partial x})v(E)\right)$

small signal assumption

$E = E_0 + E_1\exp(j\omega t)$

$I_{tot} = I_{tot0} + I_{tot1}\exp(j\omega t)$

$E_1(x) = \dfrac{I_{tot1}}{j\varepsilon A v_0(\omega/v_0 - jqN_D/\varepsilon v_0\,\partial v/\partial E)}\cdot$
$\{1 - \exp(-j\omega x/v_0)\exp(-qN_D/\varepsilon v_0\,\partial v/\partial E)\}$

$\dfrac{\partial v}{\partial E} > 0$ damping

$\dfrac{\partial v}{\partial E} = 0$ no attenuation

$\dfrac{\partial v}{\partial E} < 0$ growing up

Transit-time device

- injection region
- drift region

$I_{rf} = 2 I_0 \sin(\Theta_d/2)/(\Theta_d/2)$ Fourier component

$I_0 = I_{max} \Theta_d/(2\pi)$ dc-current

$P_{rf} = U_{rf} I_0 \dfrac{\sin(\Theta_d/2)}{\Theta_d/2} \sin(\Theta_i + \Theta_d/2)$

$\Theta_i = \omega t_i$ injection angle
$\Theta_d = \omega t_d$ drift angle

generation of output power : $\pi < \Theta_i + \Theta_d/2 < 2\pi$
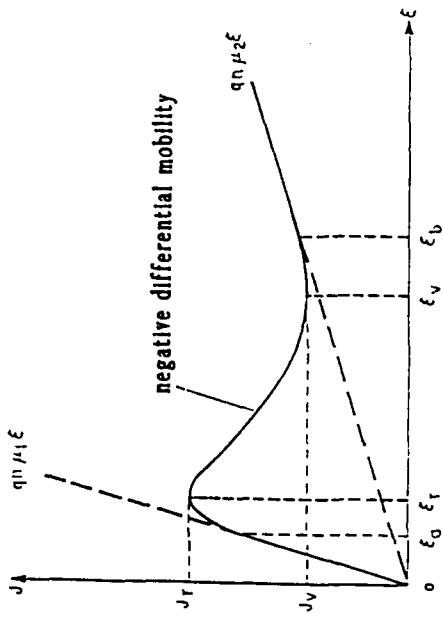
Current density versus electric field of a two-valley semiconductor

Labels: $q n \mu_1 \xi$, negative differential mobility, $q n \mu_2 \xi$, J_r, J_v, ε_a, ε_r, ε_v, ε_b
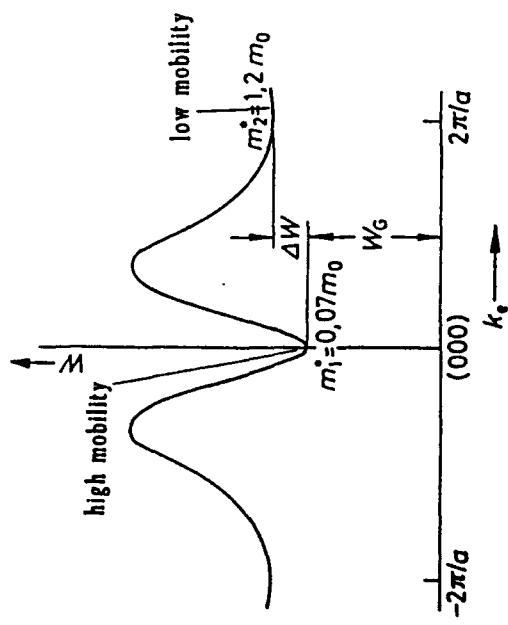
Schematical structure of a Gunn element

Labels: contact, n^+, n, n^+, contact, $-$, $+$

Energy band structure of GaAs

Labels: low mobility, $m_2^* \doteq 1{,}2\, m_0$, high mobility, $m_1^* = 0{,}07\, m_0$, ΔW, W_G, M, k_e, (000), $2\pi/a$, $-2\pi/a$

Noise of Gunn elements

- primary noise

- diffusion   (statistical velocity fluctuations of drifting carriers)
- modulation   (low frequency fluctuations of dc-current, temperature)

- up-conversion via the device impedance

Structure and field distribution of a
Read IMPATT
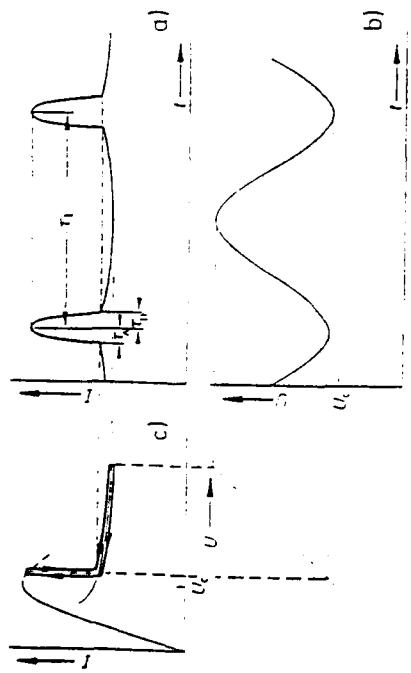
Current and voltage of a Gunn element in transit-time mode

Limitation of output power of Gunn elements

- The negative mobility vanishes if the oscillation periode reaches the scattering time

- The n^+ cathode injects "cold" electrons which must gain energy before they can transfer.
  A "dead" zone results which occupies a large fraction of the total n-zone

Improvement

- Injection of "hot" electrons

- GaAlAs/GaAs launcher
- planar doped barriers
- camel diodes
- tunnel injection
- reverse biased Schottky barriers
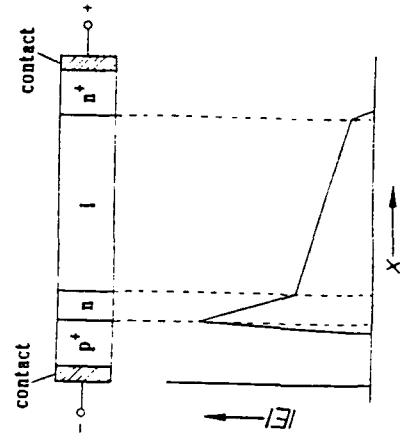
Negative resistance of IMPATT diode

$$G = \alpha v_s (J_n + J_p)$$

$$\frac{\partial J_n}{\partial t} = - v_s \frac{\partial J_n}{\partial t} + \alpha v_s (J_n + J_p)$$

$$\frac{\partial J_p}{\partial t} = v_s \frac{\partial J_p}{\partial t} + \alpha v_s (J_n + J_p)$$

$$J_c = J_n + J_p$$

$$\boxed{\frac{\tau_a}{2} \frac{dJ_c}{dt} = J_c \left(\int_0^{l_a} \alpha \, dx - 1 \right) + J_s \qquad \text{Read equation}}$$

$$\tau_a = l_a / v_s$$

J_s : saturation current density

$$R_D = \frac{1}{\omega C_d} \frac{(1 - \cos \theta_d)/\theta_d}{(1 - \omega^2/\omega_a^2)}$$

$$\omega_a = \sqrt{2\alpha' v_s J_0 / \varepsilon} \qquad J_0 : \text{dc current density}, \quad \alpha' = d\alpha/dE$$
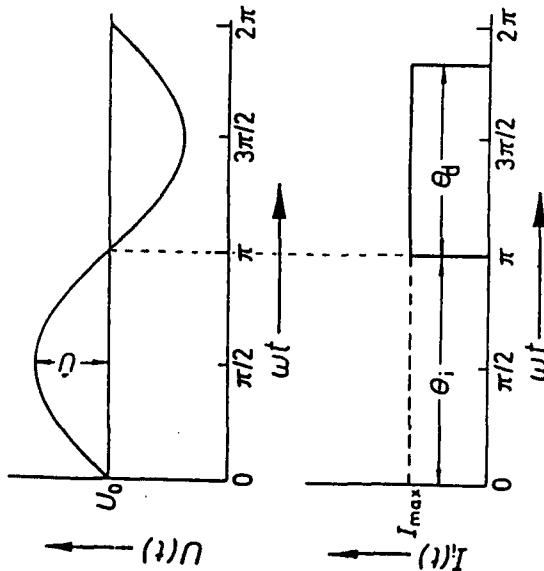
Rf-voltage and induced current of a Read IMPATT diode

IMPATT diode

$$\theta_i = \pi$$

$$P_{rf} = - U_{rf} I_0 \frac{(1 - \cos \theta_d)}{\theta_d} (1 - l_a/w)$$

l_a : length of the avalanche zone
w : total width

$$\theta_{d_{opt}} = 3\pi/4$$

optimum drift length: $l_d = \frac{\theta_d v_s}{2\pi f} = \frac{3}{8} \frac{v_s}{f}$

Limitation of output power of IMPATT diodes

- l_a/w increases with increasing doping concentration
- The fraction of the "dead" space increases with decreasing avalanche zone width
- For high frequencies the generation of carriers is reduced because of the finite intrinsic avalanche response time
- With increasing electric field more carriers are generated by tunnel effect
- The saturation drift velocity reduces with increasing electric field (ω_a is reduced)
- The ionization rates saturate with increasing electric field (ω_a is reduced)
- The thermal conductivity limits the maximum input power
  Si: 1.45 W/cmK, InP: 0.68 W/cmK, GaAs: 0.44 W/cmK

Noise of IMPATT diodes

- The IMPATT mode exhibits an intrinsically high noise level
- The noise power is mainly generated by the statistical nature of the avalanche process
- The noise behaviour is dependent on
  semiconductor material
  diode doping profile
  bias conditions
  rf-signal levels
  parametric interactions
- Generally GaAs IMPATT diode are less noisy than SI devices
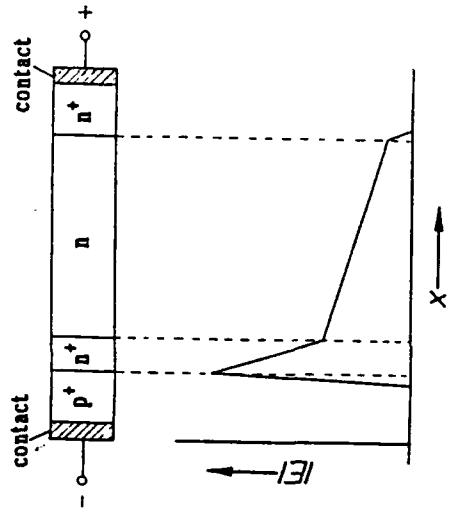
Structure and field distribution of a TUNNETT diode
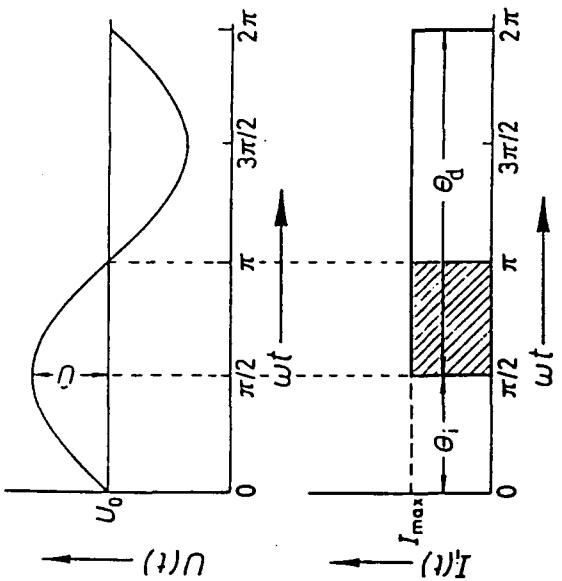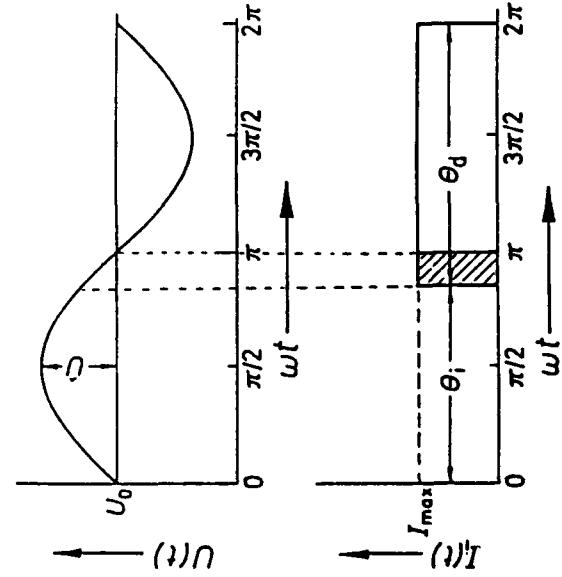
Rf-voltage and induced current of a TUNNETT diode

Rf-voltage and induced current of a TUNNETT diode

Structure and field distribution of a BARITT diode

Rf-voltage and induced current of a BARITT diode
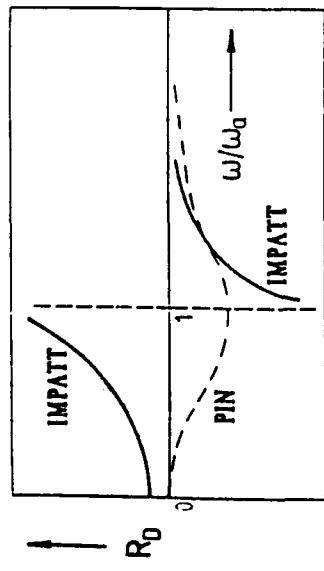
Real part of IMPATT and PIN-avalanche diode impedance versus frequency

PIN-avalanche diode

- Negative resistance not only above avalanche frequency but also below
- Broadband negative resistance
- Advantage for pulsed diodes : relatively large negative resistance at $\omega = \omega_a$
- The reactance is zero at $\omega = \omega_a$. No large capacitive currents which limit output power

Structure and field distribution of a PIN-avalanche diode

Rf-voltage and induced current of a PIN-avalanche diode

TUNNETT diode

$$\Theta_l = \pi/2$$

$$P_{rf} = u_{rf} I_0 \frac{\sin \Theta_d}{\Theta_d} (l - l_a/w)$$

l_a : length of the generation zone
w : total width

$$\Theta_{d_{opt}} = 3\pi/2$$

optimum drift length : $l_d = \frac{\Theta_d v_s}{2\pi f} = \frac{3}{4} \frac{v_s}{f}$
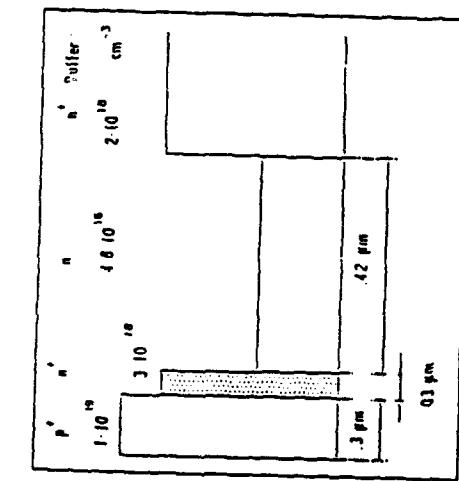
Doping profile of a W-band TUNNETT diode

BARITT diode

$\theta_i = \pi/2$

$P_{rf} = U_{rf} I_0 \frac{\sin\theta_d}{\theta_d}$

$\theta_{d_{opt}} = 3\pi/2$

optimum length: $l_d = \frac{\theta_d v_s}{2\pi f} = \frac{3}{4} \frac{v_s}{f}$
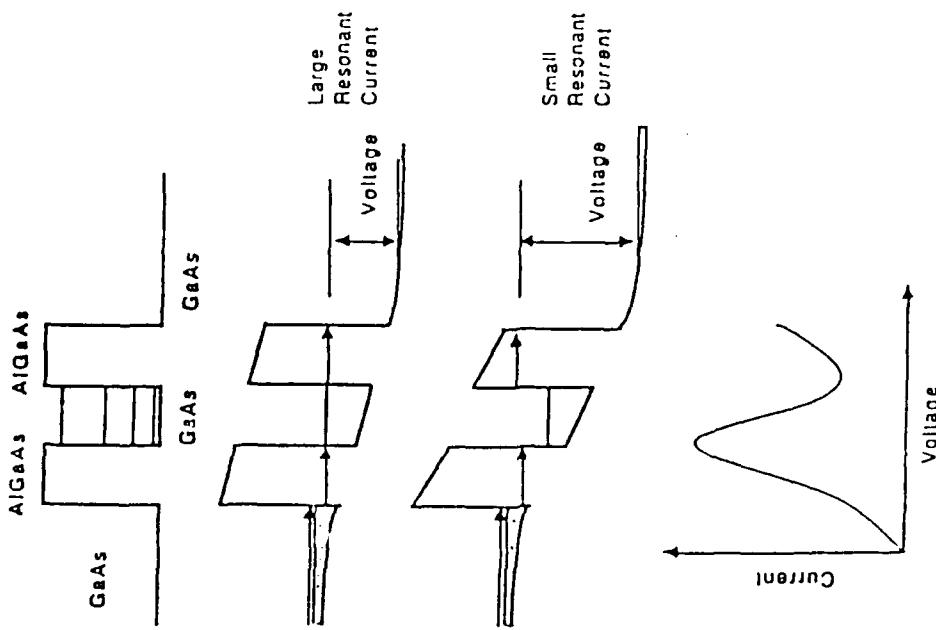
Noise of BARITT diodes

- diffusion noise of drifting carriers
- shot noise due to the injection mechanism
- low frequency noise (up-conversion)

- Phase noise as low as for Gunn elements

- Application as self-oscillating mixer
  60 GHz : minimum detectable signal ≈ -160 dBm
  conversion gain ≈ 30 dB
  (15 to 20 dB better than for Gunn elements)

GaAs

AlGaAs   AlGaAs

GaAs   GaAs

GaAs

Large Resonant Current

Voltage

Small Resonant Current

Voltage

Current

Voltage

Energy band structure of a quantum-well resonant tunneling device

electron energy →

Undoped
AlGaAs →

n+ -GaAs

undoped
GaAs

undoped GaAs

W

n+ -GaAs

E_F
E_C

E_1

E_C
E_F
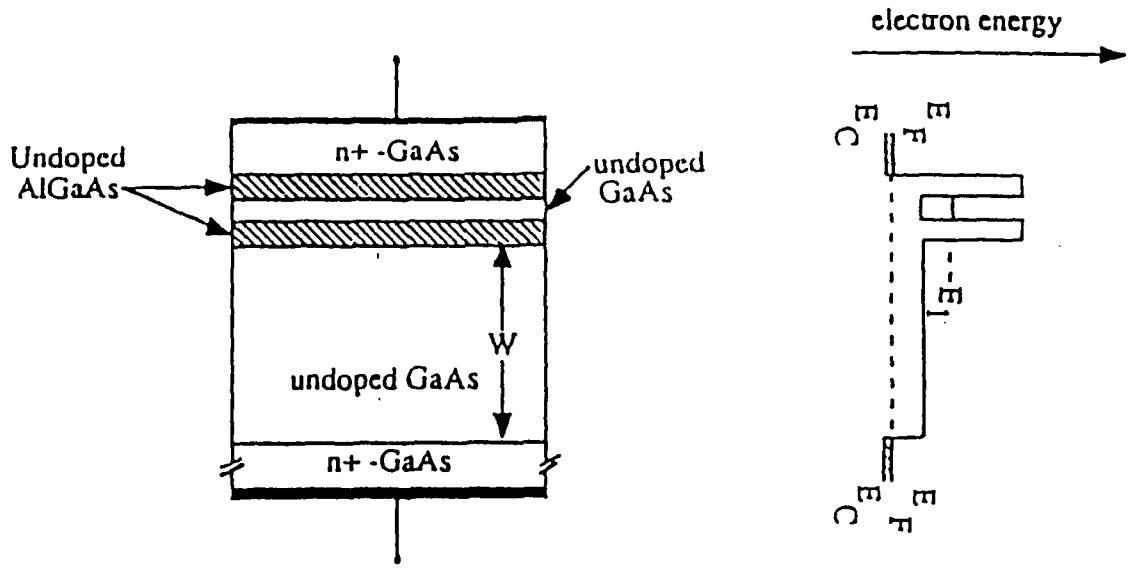
Structure and energy band diagram of a GaAs/GaAlAs QWITT diode

$I(t)$ →

$I_i(t)$ →

$U(t)$ →

I_{max}

I_{max}

U_0

ωt

ωt

ωt

$\hat{U}$

Θ_i

Θ_i

Θ_d

Θ_d

0 $\pi/2$ π $3\pi/2$ 2π

Rf-voltage and induced current of a QWITT diode ($\Theta_1 = \pi/2$ and $\Theta_1 = 3\pi/2$)
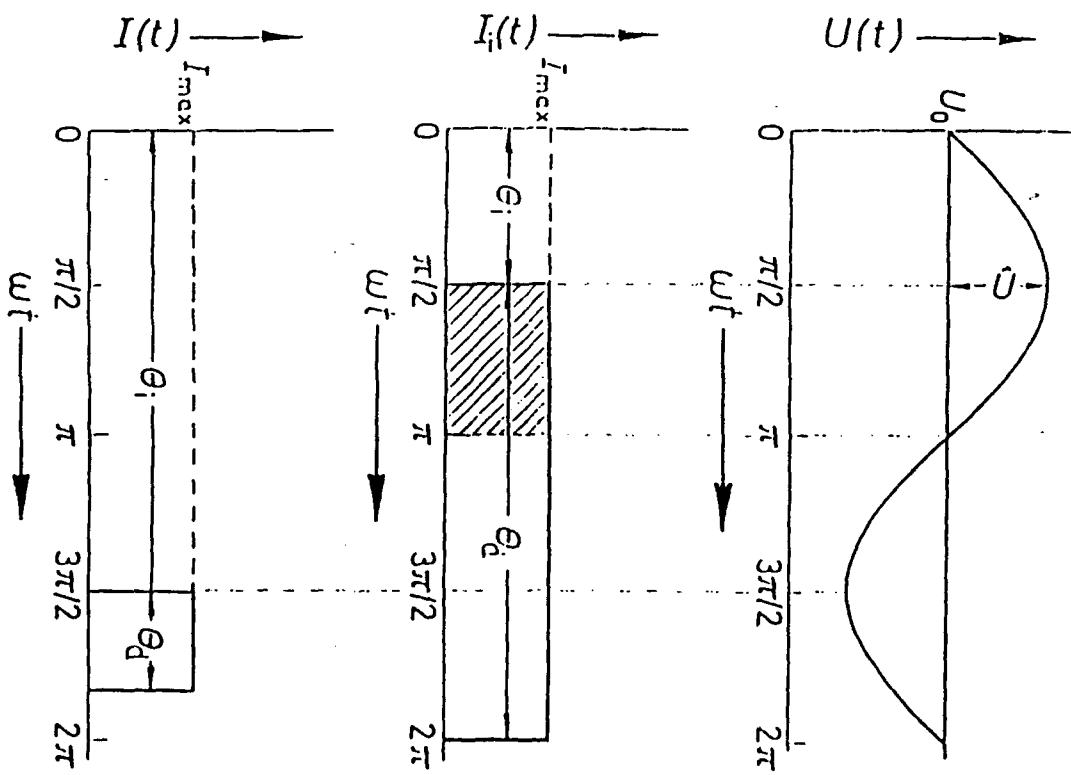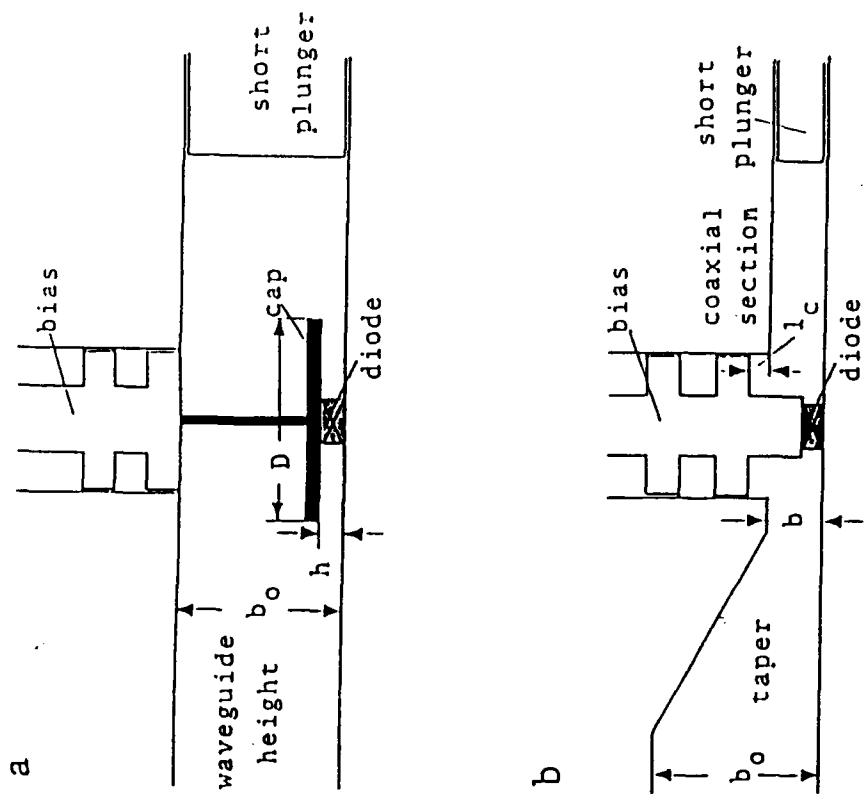
Full and reduced waveguide
height resonator

a

waveguide height b_o

short plunger

bias

cap

diode

D

h

b

coaxial section

short plunger

bias

l_c

diode

b

taper

b_o

Encapsulated W-band GaAs IMPATT
diode

100MH   20kV   02   002   8

26
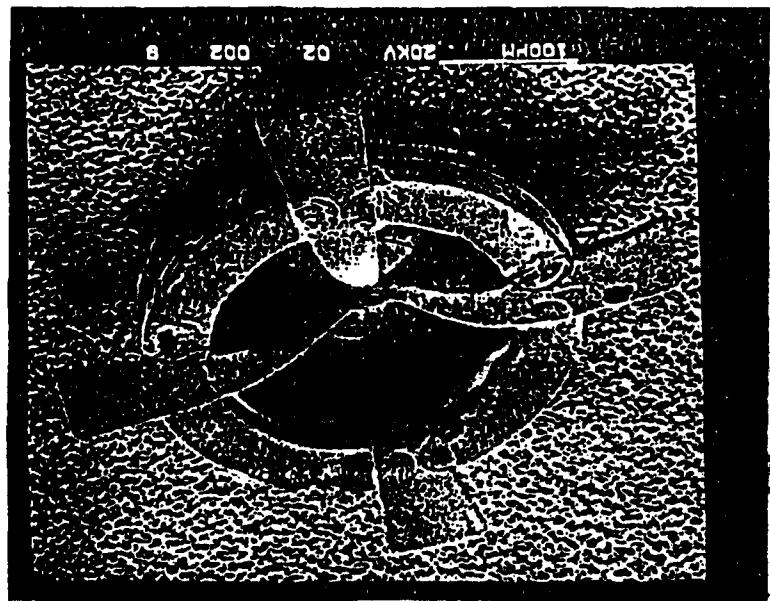
Device fabrication

Si, GaAs, InP for mm-wave devices

- Molecular beam epitaxy
  (important for elevated frequencies)
- Photoresist technology
- Etching processes (wet, dry)
- Metallization (evaporation, sputtering, plating)

Reduction of losses

- low ohmic contacts
- substrate thickness : 1 - 3 μm
- low heat flow resistance (diamond)
- reduction of parasitic capacitance and inductance
- matching to the resonator impedance
  94 GHz : C < 0.07 pF, L < 15 pH
  quartz ring height < 75 μm
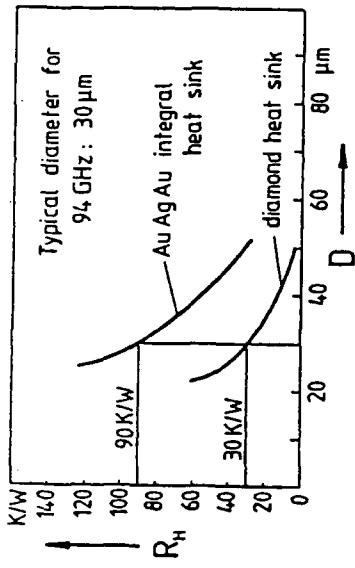
Heat flow resistance of 94 GHz IMPATT diodes versus device diameter

K/W
140 —
120 —
100 —
90 K/W
80 —
60 —
40 —
30 K/W
20 —
0 —

R_H

Typical diameter for 94 GHz: 30 μm

Au Ag Au integral heat sink

diamond heat sink

D ——▶

20   40   60   80 μm

b)

(0.4 μm) Au (~2 μm)
AuGe-Ni-0.2 μm)
n⁺-GaAs (1-2 μm)
n⁻-GaAs (0.3-0.4 μm)
p⁺-GaAs (0.3-0.5 μm)
Ti (~0.03 μm)
Pt (~0.1 μm)
Au (0.4~0.5 μm)

diamond

a)

Au
Ag
Au
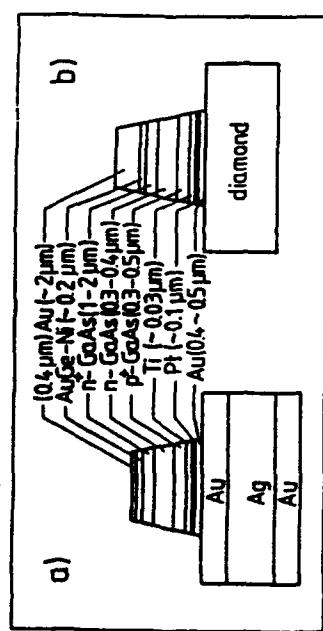
GaAs IMPATT diode with Au/Ag/Au and diamond heat sink

State-of-the-art performance of solid-state power devices (after Shih and Kuno)

Si BJT

MESFET

GaAs MMIC

IMPATT

GUNN

OUTPUT POWER (W)

10^3

10^2

10

1

10^{-1}

10^{-2}

10^{-3}

FREQUENCY (GHz)

10

100
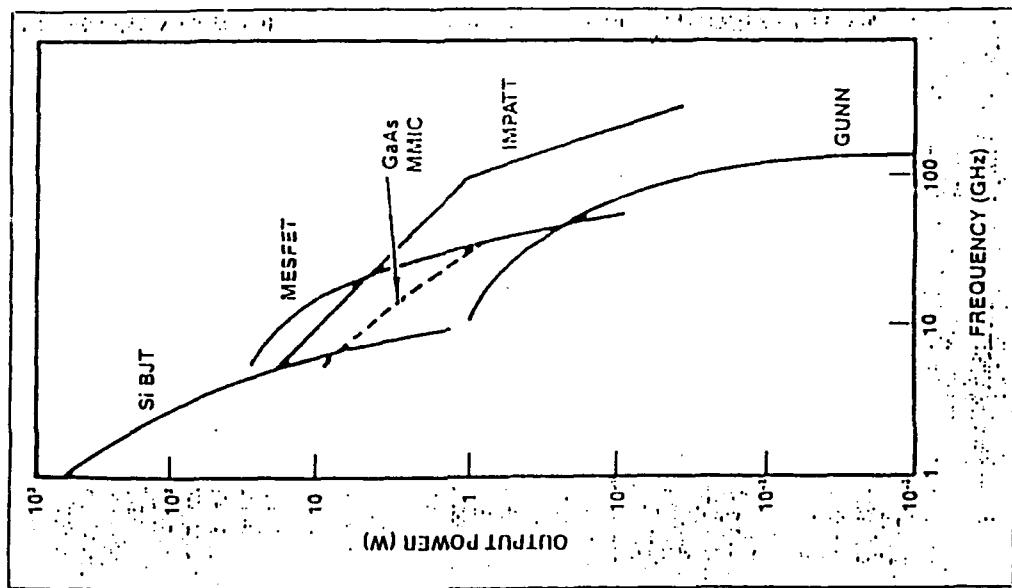
Photograph of a $\lambda/4$ strip-line resonator on semiinsulating GaAs substrate

Impatt diode

resonator

substrate

heat sink

output line

dc bias
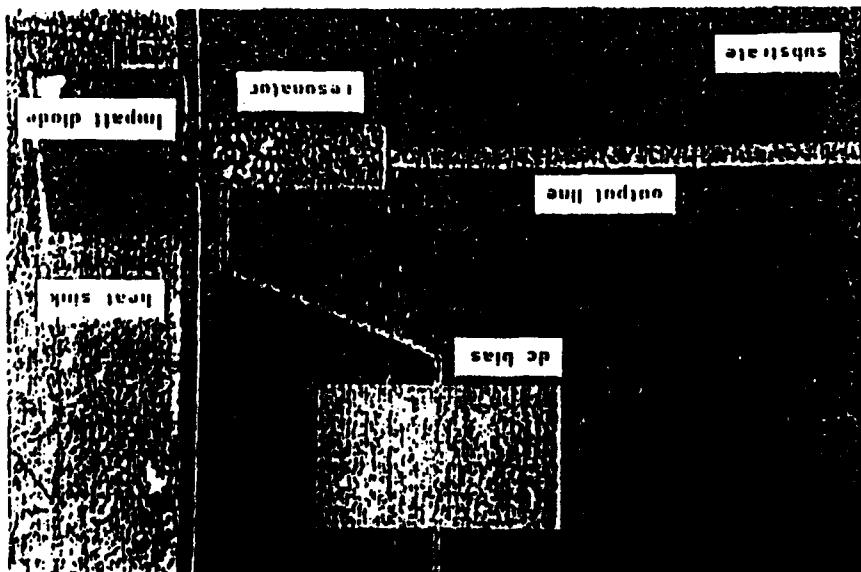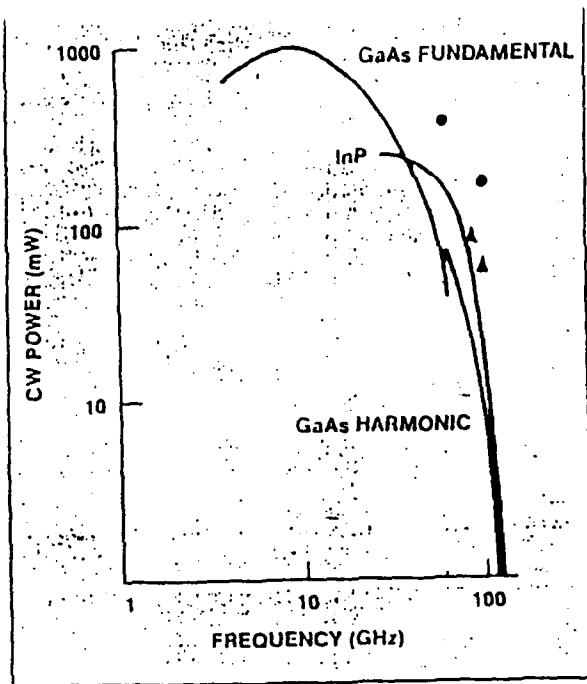
Cw-output power of Gunn elements
(▲ GaAlAs/GaAs launcher
● InP current limiting cathode)
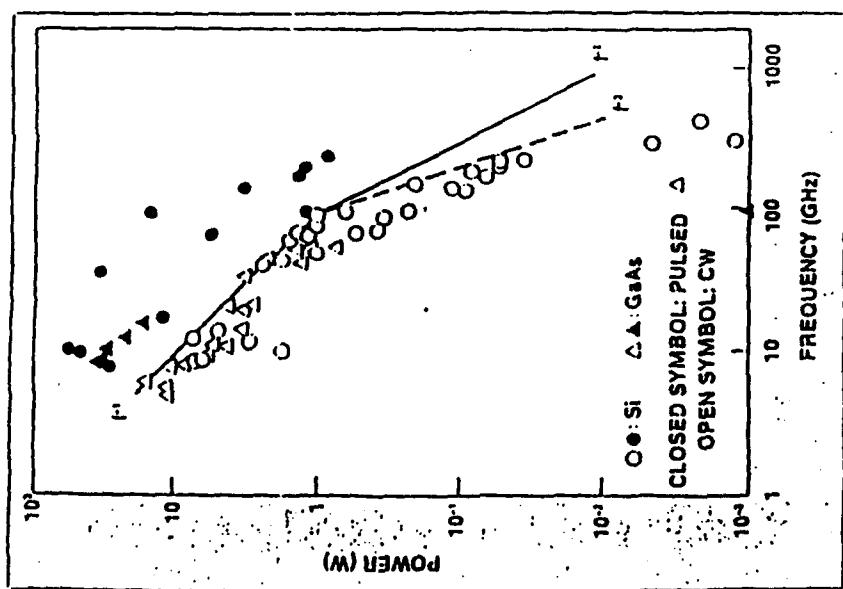
State-of-the-art performance
of Si and GaAs IMPATT diodes
After Chin and Kuno)

Maximum efficiency

Impatt   20 %   f < 40 GHz   for GaAs devices
         10 %   f > 40 GHz

         10 - 12 %   for silicon devices

Gunn   2 -5 %   for GaAs devices
         5.7 % at 94 GHz   for InP devices
         (12 % at 60 GHz, pulsed,   for InP

TUNNETT 2 %   at 94 GHz

BARRITT 0.4 %   at 60 GHz

MITATT   0.5 %   at 150 GHz
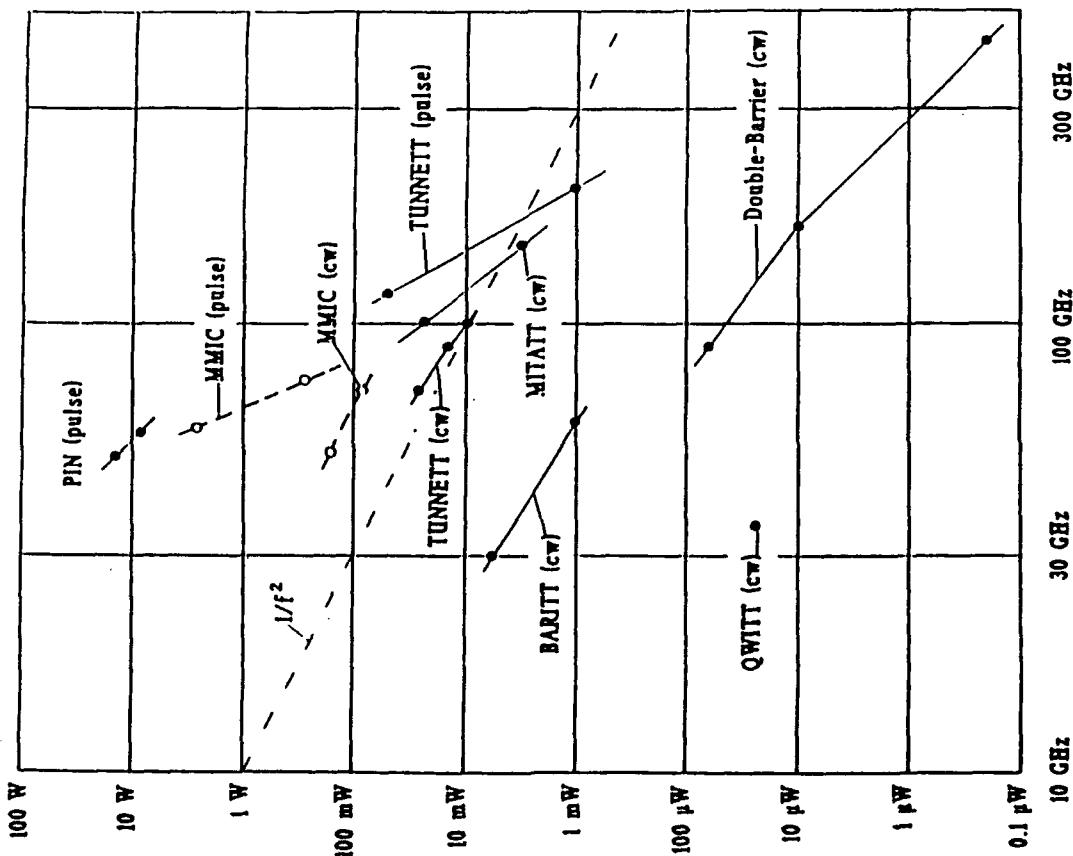
Cw and pulsed output power
of different mm-wave devices

Noise measure

$$M = \frac{F - 1}{1 - 1/G} = \frac{\overline{u_n^2}/B}{4kT_0|R_D|}$$

F   : Noise figure
$\underline{G}$: amplification
$\overline{u_n^2}$: primary noise voltage source
B   : bandwidth
k   : Boltzmann constant
T_0 : room temperature
R_D : negative diode resistance

$$M = \frac{P}{kT_0 B} \frac{\Delta f_{rms}^2 Q_{ex}^2}{f_0^2}$$

94 GHz : $M_{TUNNETT} \approx 18$ dB

$M_{IMPATT} \approx 21$ dB

P   : rf power
Δf_{rms} : rms frequency deviation
f_0 : oscillation frequency
Q_{ex} : external quality factor
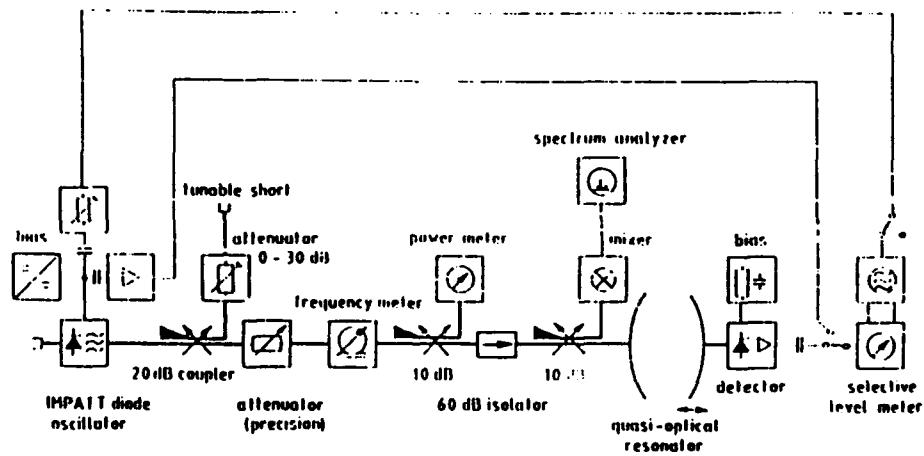
Noise measurement set-up

TECHNOLOGY DEVELOPMENTS TOWARDS NANOMETRIC STRUCTURES

H. L. Hartnagel

University of Darmstadt • Germany

Technology Developments
towards
Nanometric Structures

H. L. Hartnagel
Techn. University Darmstadt
Inst. f. Hochfrequenztechnik
Merckstr. 25
61 Darmstadt
Germany

) Basic Considerations towards Nanometric
   Devices

:) Selective Epitaxy

;) E-beam Lithography and soft
      Electrolytic Processes:
         Anodic, Etching and Cathodic
            metal deponition.

, Local-Structure Deposition by point-like
      Electrode Current scanned by
      Piezo-Actuators.

) Evaluation of Nanometric Surfaces

) Outlook

Basic Considerations towards
Nanometric Devices

— Dimensional Reduction

a) hot-electron transport ——▶ High Speeds

b) high packing densities for highly complex
   signal processing

c) Impedance optimization for capacitances
   based on optimul space-charge layers
   for mm waves:

$$|z_c| = \frac{1}{\omega C_j}$$

d) minimization of heat loss for current
   crowding along emitter periphery of
   power transistors for mm waves

— Increase of Surface to Volume Ratio

a) no damage acceptable leading to traps

b) space-charge layer minimization to obtain
   carrier-density independent, controllable
   conditions
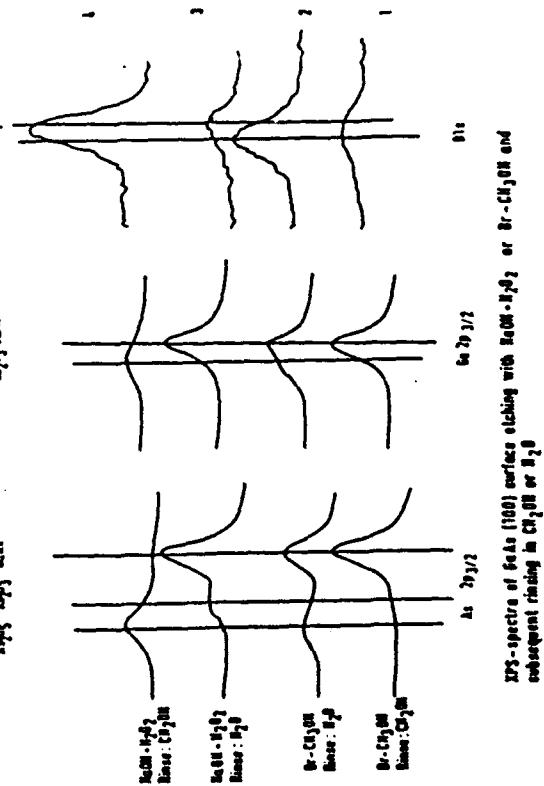
XPS-spectra of GaAs(100) surfaces after different surface treatments. The As 2p3/2, As3d, Ga 2p3/2, Ga 3d and O 1s core level peaks are shown

XPS-spectra of GaAs (100) surface etching with NaOH·H2O2 or Br-CH3OH and subsequent rinsing in CH3OH or H2O

ISS spectra of GaAs
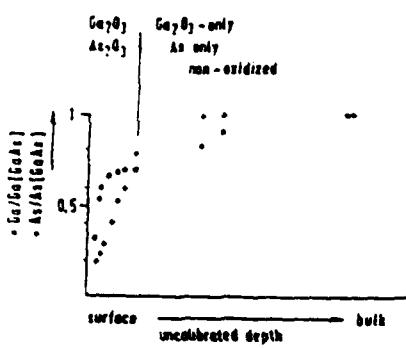
Depth profile of Ga and As by detailed ^{20}Ne-ISS analysis
Ga|GaAs| : concentration of Ga in GaAs-compound
As|GaAs| : concentration of As in GaAs-compound
Ga : concentration of As after the given etch step
As : concentration of As after the given etch step
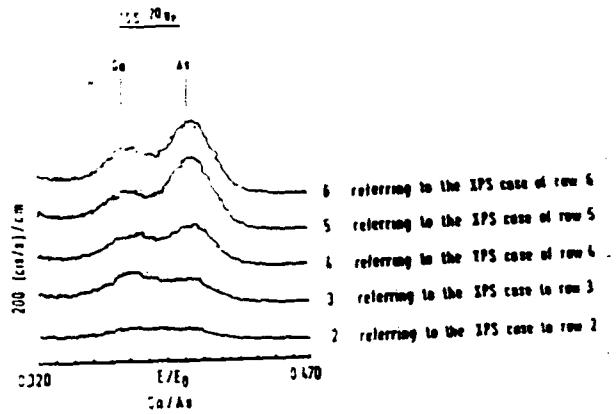
^{20}Ne-ISS spectra

XPS-spectra
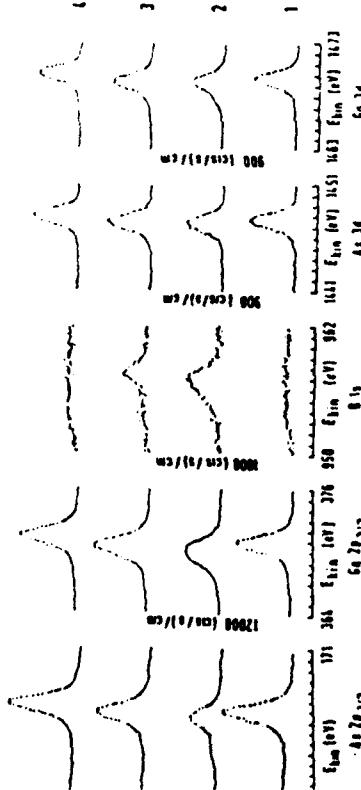
row 1 identical with Fig. 3 row 6
row 2 after reoxidation in air for 30 min
row 3 after short ionetching
row 4 after heavy Ar etching similar to that before spectrum given by Fig 3 row 6

XPS Al$_{K\alpha}$ hν=1486.7
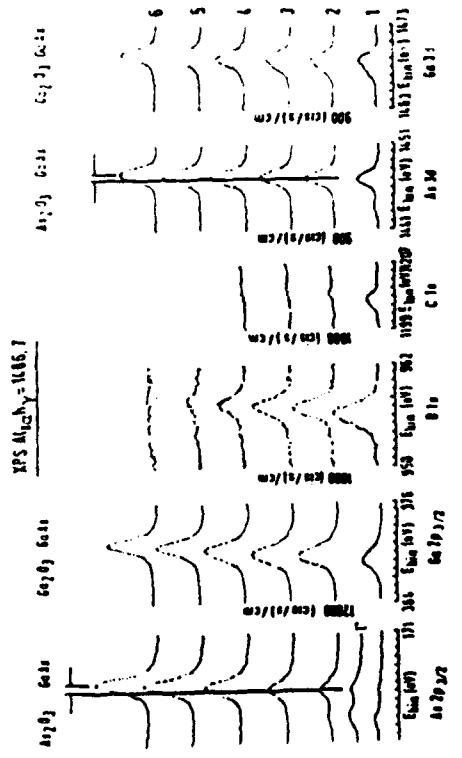
XPS-spectra
row 1 original GaAs-sample as inserted into the vacuum system
row 2 to 5 subsequent etching with Ar$^+$ of different intensities and duration
row 6 last etch step when Ar$^+$ was used (Ar$^+$ is much heavier than Ne$^+$)
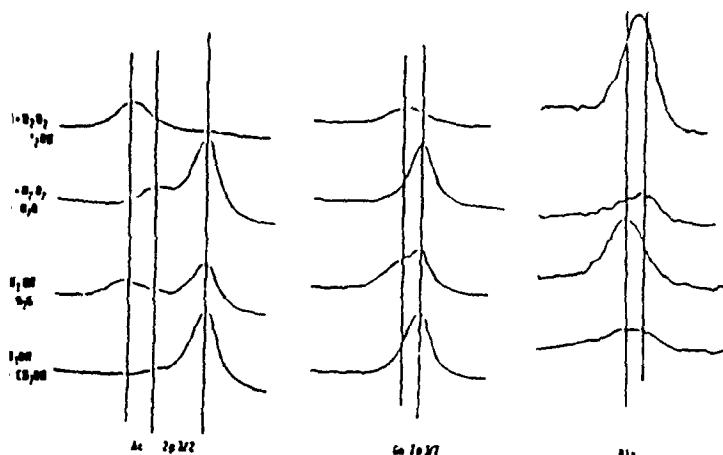
XPS-spectra of GaAs(100) surface etching with NaOH + H_2O_2 or Br-CH_3OH and subsequent rinsing in CH_3OH or H_2O

XPS-SPECTRA OF DIFFERENTLY TREATED (100)-GaAs-SURFACES
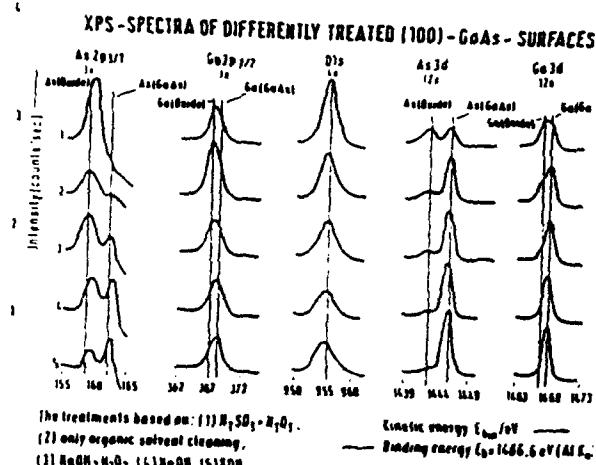
The treatments based on: (1) $H_2SO_4 + H_2O_2$.
(2) only organic solvent cleaning.
(3) NaOH + H_2O_2. (4) NaOH. (5) EtOH

Kinetic energy E_{kin}/eV ———
Binding energy $E_b = 1486.6$ eV (Al K_α)

XPS-SPECTRA OF SAMPLES TREATED BY NH_4OH (25%) : H_2O_2(30%): H_2O (2:1:300)
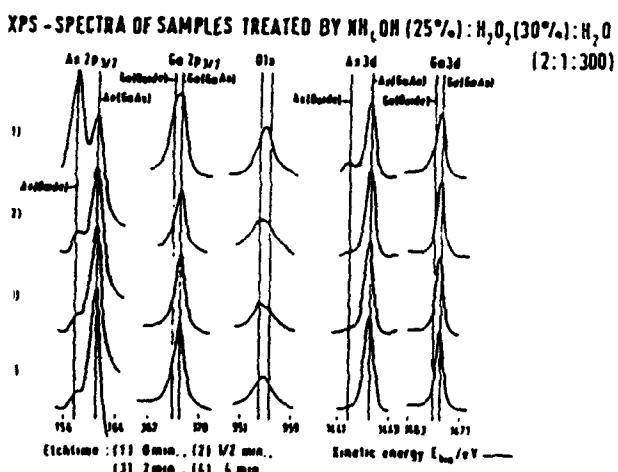
Etchtime : (1) 6min., (2) 1/2 min.,
(3) 2min., (4) 4 min.

Kinetic energy E_{kin}/eV ———

Selective Epitaxy

Starting with high-temperature capable amorphous mask

Growth required with good quality on side walls

Same growth rate as with volume case to allow fabrication of lateral acces structures
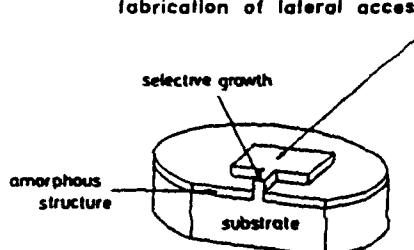
selective growth

amorphous structure

substrate

E–Beam Lithography and Soft Electrolytic
Processes:
  Anodic Etching and
  Cathodic Metal Deposition

1) nanometric structurization by electron–beam
   illumination

2) gas–plasma etching through E–beam resist
   pattern
   a) steep structures possible

   b) but ion bombardment
      minimized only by protective
      grids or plasma potential
      control

3) anodic wet etching through resist pattern
   a) side–steepness by combined
      adjustment of bias and opt.
      illumination

   b) passage through verious epitaxial
      and doping layers by optical
      hole generation as required for
      anodic process

Ion–Beam Lithography

H^+, 60 KeV into PMMA
spread: o.1 μm
no proximity effect
no backscattering on Si

-I6.

E–Beam Lithography

pos. resist

PMMA polymethyl metacrylate
PBS polybutane –1 sulfone
     resulution: 1   m

neg. resist:

COP Polyglycidyl methacrylate –
         co– ethyl acrylate
     resulution due to swelling: 1 μm

limitation:

     electron scattering, 20 keV
     ($\lambda \sim 1\text{Å}$); proximity effect
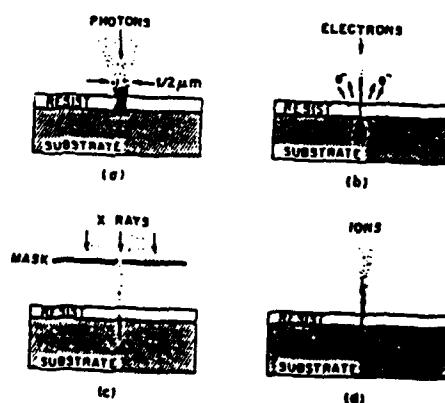
Types of advanced lithographic methods. (a) Optical lithography. (b) Electron beam lithography. (c) X–ray lithography. (d) Ion beam lithography.

ə Sub-mm-Wellen-Mischung

Continuation: Electrolytic Processes

3)   c) Selection of electrolyte:
                reaction—limited solution
                instead of diffusion
                limited one

            d) Equal solution rates
                for each of the components
                of the semiconductor

            e) pulsed—bias

4)   cathodic metal deposition through resist pattern

            a) resulting metal thickness
                proportional to local current
                density

            b) therefore illumination to increase
                current flow over layers of
                reduced conductivity

            c) sometimes side—wall metallization
                avoided by intermediate
                insulating epi layer and
                deposition in darkness

Schematischer Diodenquerschnitt

Changes of surface structure
in air after treatment b)
measured by STM
(15min between each

measurement)

:ated structures (1):

of GaAs towers

300 nm

600 nm

2.5 um
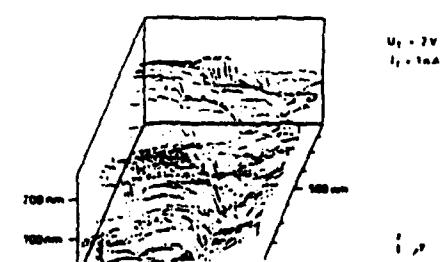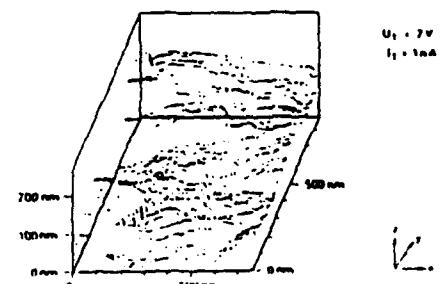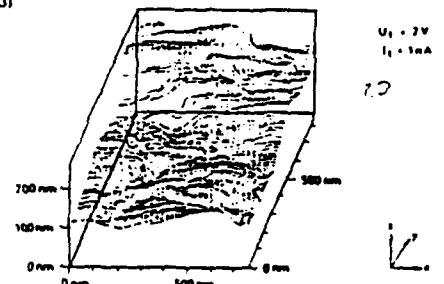
STM-Studies of GaAs(100) Surfaces after different etch treatments

alkaline etchant
(40H:H202:H20
(:1:300)

U_t . 15V
I_t . 1nA

photochemical passivation
. then 30min
(Cl:H20 1:1, UV-light)

U_t . 2V
I_t . 1nA

-passivation
. then 10min in (NH4)2S)

U_t . 15V
I_t . 1nA

SiO_2

$$SiH_4 + O_2 \longrightarrow SiO_2 + 2H_2$$
$$(450°C)$$

$Si(OC_2H_5)_4$ Tetraethylorthosilicate
= TEOS   (700°C)
(from liquid source)

$SiCl_2H_2 + 2N_2O$ (900°C)
dichlorosilane
nitrous oxide

S_iN_4

LPCVD   : 750°C
PCVD   : 300°C
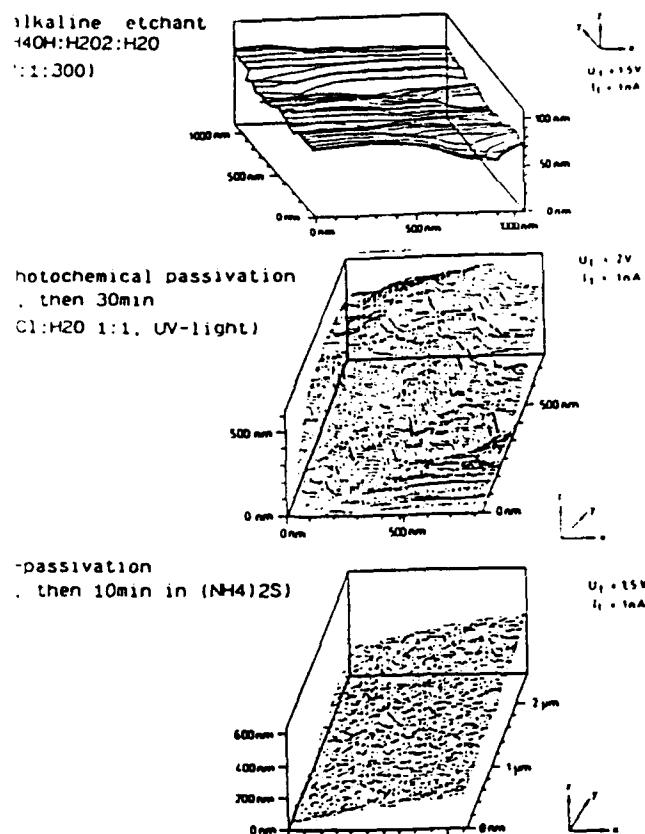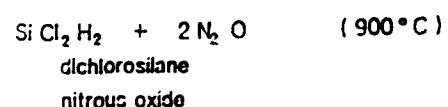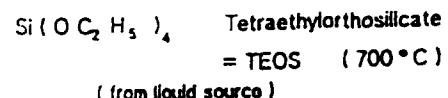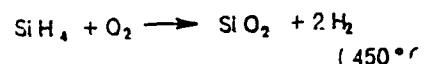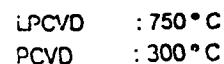
(final chip passivation against
scratch protection, moisture
barrier, sodium barrier)

$3SiCl_2H_2 + HNH_3$ (LPCVD)
$SiH_4 + NH_3$
$SiH_4 + N_2$
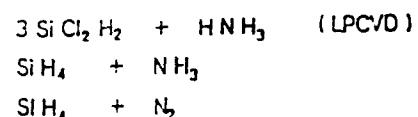
Poly Si   (larger stability than Al)

$$SiH_4 \longrightarrow Si + 2H_2 \quad (600°C)$$
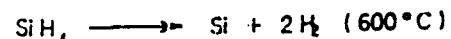
Chemical Vapour Deposition

(e.g. SiO₂ to insulate multilevel metallisation)

1) atmospheric -pressure CVD

2) low -pressure CVD

3) plasma -assisted CVD

- Hot -wall reactor
- Low -Temperature CVD for Si submicron techn.)
  (reduced thermal broadering of dopants
                              by diffusion)
  energy -enhanced CVD
  (possible: SiO₂ at 150 Å/min for 50°C
  by UV light ; in
  " Chemical Vapour Deposition "
  1984, ed.: M.D. Robinson et al
  Electrochem. Soc., Pennington, USA)

Cont.: Electrolytic Processes

4)   d) metal sequences by several
           cathodic steps as required
           for
           - ohmic contacts
           - Schottky contacts
           - good addhesion
           - high stability (diffusion
             barrier)

       e) selective deposition on
          semiconducting islands in
          semiinsulating surface in
          darkness

       f) pulsed bias
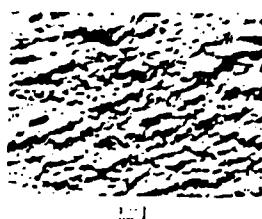
DOUBLE-REPLICA TEM PHOTOGRAPH OF
$H_2SO_3: H_2O_2$ ETCHED GaAs SURFACE
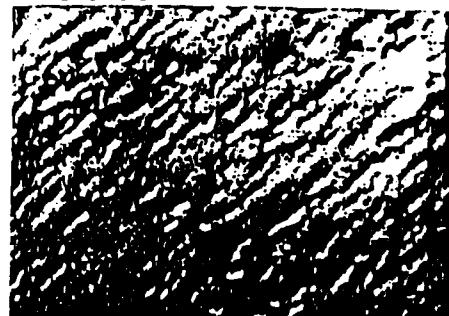
DOUBLE-REPLICA TEM PHOTOGRAPH OF Al-FILM SURFACE
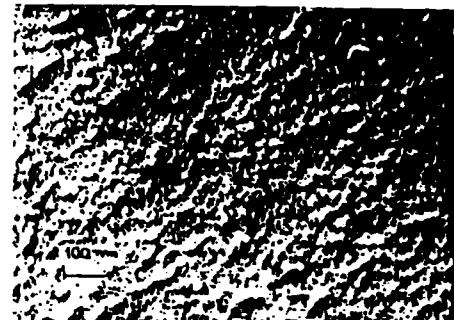
Local Structure Deposition by Point—like Electrode Current scanned by Piezo—Actuators

1) Point Current Source by

   a)   needle with insulation except on tip

       metal
       insulating
       local cathodic deposition
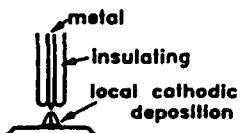
   b)   fine capillary in conducting electrolyte

       capillary filled with electrolyte
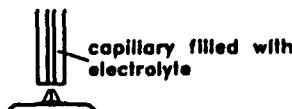
2) Pulsed Bias for optimization of results

       flow of metallic source in electrolyte
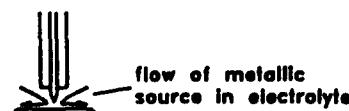
3) Position control by piezo actuators similar to Scanning Tunnelling Microscope.

4) Combination with local etching under point current source

   subsequent local   first local
   metal deposition   etching
   on ridge
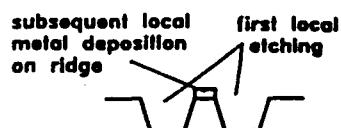
5) Surface preparation by focussed—electron (SEM) or focussed—ion illumination for local etching with point or line metallizations.

Direct writing of submicron metallisation patterns on GaAs

Development of a new process for writing submicron metallisation patterns with a tungsten-tip on n-GaAs in an elctrolytic Au-bath

Experimental set-up

experimental setup:

- pulse generator
- teflon holder
- Pt anode
- electrolytic solution
- magnetic stirrer
- photoresist- or e-beam-resist structured GaAs sample
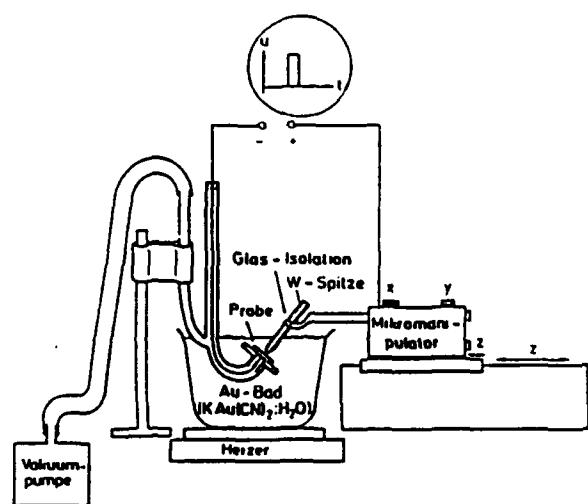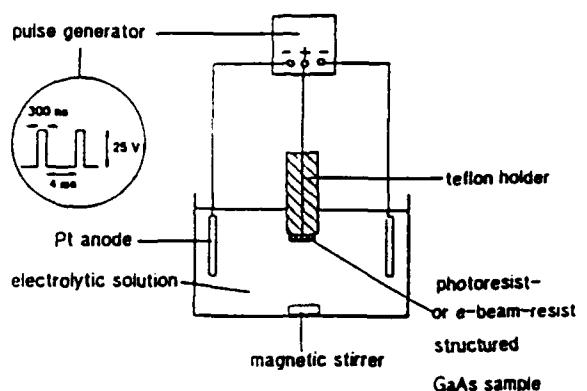
Pt-solution temperature: 40°C

etch rate: 4 nm/pulse

3.Results

-dots:

diameter from 100 nm to 3 μm

height $\frac{1}{2}$ that the diameter

8
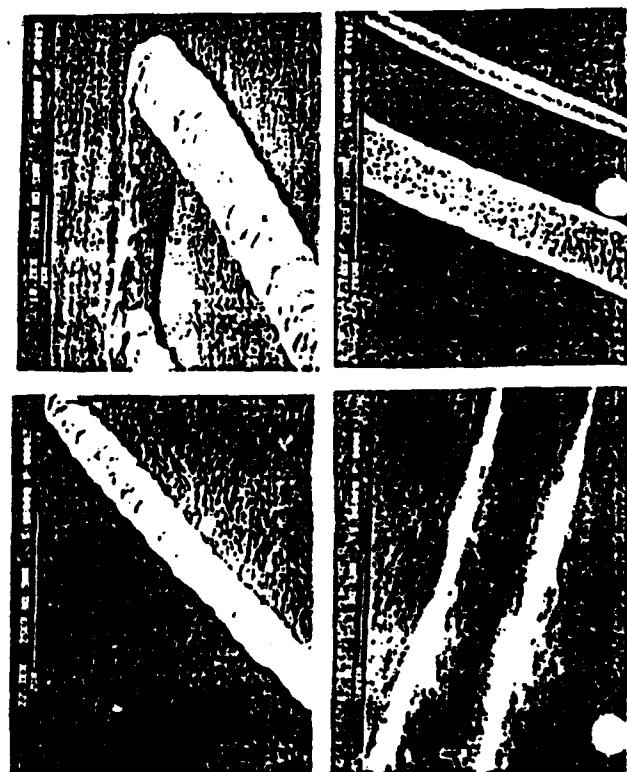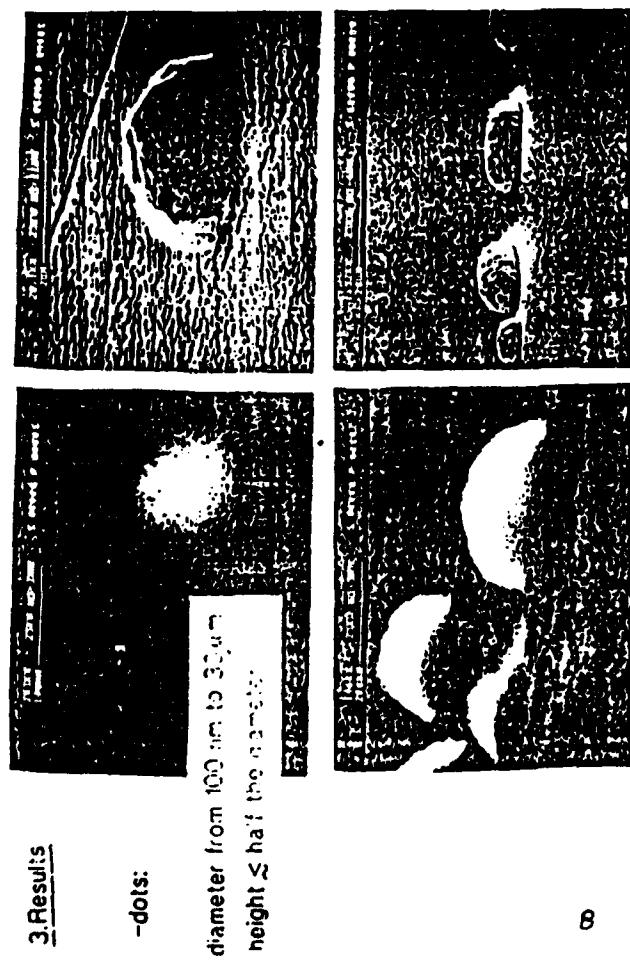
Evaluation of Nanometric Surfaces

1) STM and its derivatives in particular scanning
   tunneling luminescence

2) Electron-beam evaluation

   a) SEM

   b) Scanning beam
      cathodoluminescence

   c) TEM, particularly at angle or
      transverse; possibly with double
      replica techniques.

3) Focussed-Ion techniques
   incl. Ion Scattering Spectrosscopy

— Soft structurization
— Local epitaxy
— relevant structure characterization
        will give
— mm-wave devices and IC's
— high packing densities for complex signal
    processing with short interconnection
    delays
— extension to μm waves
— broad-band approach

9

OPTICAL CONTROL OF MILLIMETER - WAVE CIRCUITS

T. Itoh

UCLA · USA

PROGRESS OF A TUNABLE ACTIVE BANDPASS FILTER

Professor Tatsuo Itoh

Electrical Engineering Department

University of California, Los Angeles

U.S.A.

AGENDA

- Introduction
- Passive Bandpass Filter
- Active Bandpass Filter
  - coupled negative resistance method
- Tunable Active Bandpass Filter Using Varactor Diode
- Tunable Active Bandpass Filter Using MESFET as Varactor
  - optical tuning as well as electrical tuning
- Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor
  - different ways of using MESFETs as varactors
  - passive mode, active mode
- Conclusion

Introduction

- Filter in MMIC form is difficult to realize because of the low Q value for MMIC element.
  → *Active Filter*

- Direct optical control of microwave devices has been an area of growing interest.
  → *Tunable Active Bandpass Filter Using MESFETs*

- Use of both negative resistance and reactance of an active device has design potential in microwave active circuit.
  → *Tunable Active Bandpass Filter Using one MESFET*

Microstrip-line Bandpass Filter

- Passive circuit
  - loss is large, Q value is low
- Half-wavelength resonator
  - determine the frequency of each pole
- End-coupled bandpass filter
  - Chebyshev response if lossless

Multi-pole bandpass filter

One-pole bandpass filter

Active Bandpass Filter

- Comparison of passive filter and active filter
- The Q value is increased by coupled negative resistance method
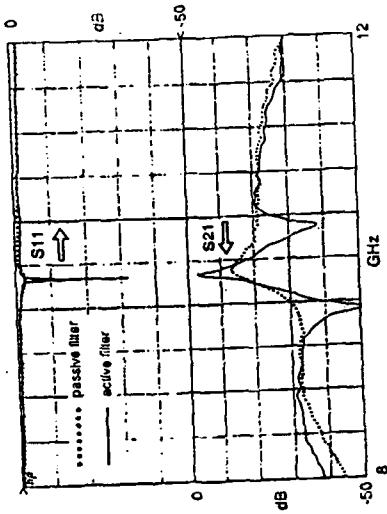
Active Bandpass Filter

- Multi-pole configuration
- Coupled negative resistance in each tank circuit
- Chebyshev response

Active Bandpass Filter

Simulation Result of Negative Resistance Using NE71083

Re [Z]
Im [Z]

Active Bandpass Filter

- Coupled negative resistance –
  compensate the loss of resonant tank, increase the Q value
- Fixed center frequency of passband

quarter-wavelength coupled line

INPUT

OUTPUT

MESFET

Substrate: RT/duroid® ULTRALAM
dielectric constant 2.55
Substrate thickness: 30mil
Conductor thickness: 1.4mil
GaAs MESFET: NE71083

D : drain
G : gate
S : source
R : termination resistor

Tunable Active Bandpass Filter Using Varactor Diode

- 500 MHz tuning range
- MESFET bias voltage changs to compensate the different Q value of the tank
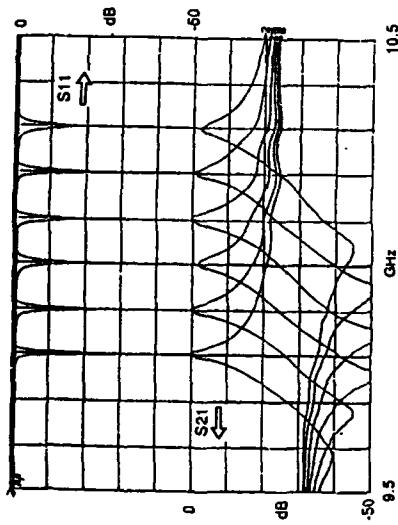
IEEE Trans. on MTT, MTT-38, December 1990

Tunable Active Bandpass Filter Using

- Gate-to-source capacitance is a function of bias v
  - Used to replace the varactor diode

semi-insulating layer

Tunable Active Bandpass Filter Using Varactor Diode

The function of varactor diode :
  change the effective electrical length of half-wavelength resonant tank, thus change the center frequency of the passband

Two quarter-wavelength sections are connected by the varactor diode

$\frac{1}{4}\lambda_g$ $\frac{1}{4}\lambda_g$

varactor diode

$C=C(V)$, variable capacitance

Tunable Active Bandpass Filter Using Varactor Diode

INPUT

OUTPUT

Varactor

MESFET

Substrate: RT/duroid® ULTRALAM
  dielectric constant 2.55
Substrate thickness: 30mil
Conductor thickness: 1.4mil
GaAs MESFET: NE71083

D : drain
G : gate
S : source
R : termination resistor

Tunable Active Bandpass Filter as Varactor

• Gate-to-source capacitance is a function of bias voltage
 – used to replace the varactor diode

variable capacitance

depletion region

semi-insulating layer

D : drain
G : gate
S : source
R : termination resistor
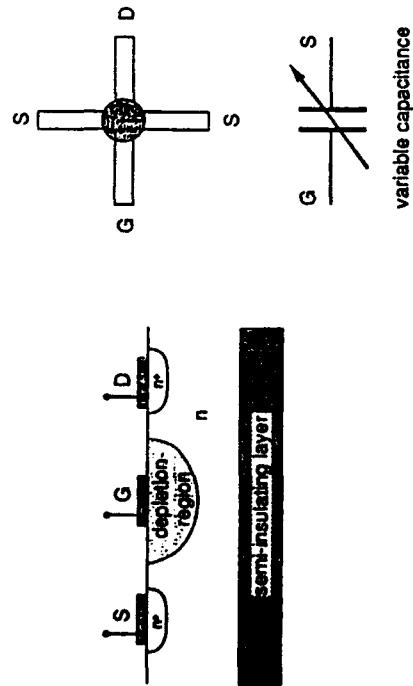
Tunable Active Bandpass Filter Using MESFET as Varactor

• electrical-tuning result
• 75 MHz tuning range without compensation of Q value

V_{GC} -1.94v -2.17 -2.29 -2.40 -2.48 -2.58 -2.68 -2.86
-1.81

S_{11} (dB)
S_{21} (dB)

Frequency f (GHz)
9.65 9.70 9.75
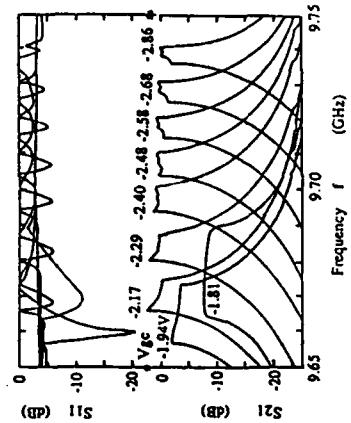
IEEE Microwave and Guided Wave Letters, May 1991

Tunable Active Bandpass Filter Using MESFET as Varactor

• Both devices in this circuit are MESFETs – advantage

INPUT

OUTPUT

MESFET (varactor)

MESFET (negative resistance)

$-V_g$

V_g

V_d

Substrate: RT/duroide ULTRALAM
dielectric constant 2.55
Substrate thickness: 30mil
Conductor thickness: 1.4mil
GaAs MESFET: NET2004
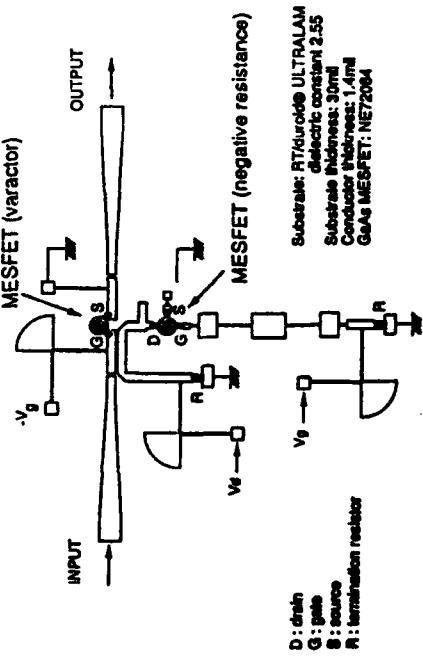
Tunable Active Bandpass Filter Using MESFET as Varactor

• Optical control of solid state device

light incidence

788nm (red light) laser diode

MESFET without package cover

charge generation occured
in the depletion region
–> change the capacitance

depletion region

semi-insulating layer

Tunable Active Bandpass Filter Using MESFET as Varactor

- optical-tuning result
- 225 MHz tuning range
- MESFET (negative resistance) bias voltage changs to compensate the different Q value of the tank

I_L (mA)	Laser Power (mW)
82	15
75	12.5
61	7.7
47	3
0	0

IEEE MTT-S, June 1991

MESFET
NET20843

Tunable Active Bandpass Filter Using MESFET as Varactor

- Optical Tuning
  - frequency shift v.s. laser power

$\Delta f \propto \sqrt{P_L}$

Tunable Active Bandpass Filter Using MESFET as Varactor

Cgs of the MESFET as a function of Vgs at two different levels of light illumination: PL = 0 mW and PL = 15 mW (PL is the output power of a semiconductor laser diode, Sharp LT021MC0)
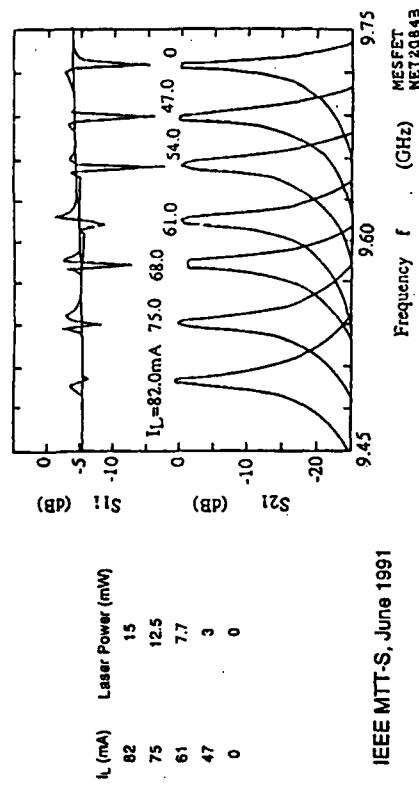
Tunable Active Bandpass Filter Using MESFET as Varactor

Cgs and Ggs as a function of the output power from a semiconductor laser diode(Sharp LT021MC0).
Bias voltage (Vgs) = -2.21V
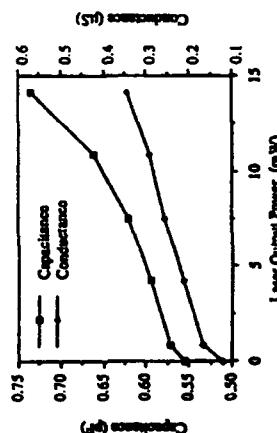
Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- Reactance-tuning circuit

open-circuited

V_d

D S

G

MESFET

input

V_g

- The MESFET is biased in the passive mode
- The impedance at the input port is nearly reactance
- If resistance presents at the input port – Q value is affected

Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Passive Mode

- Simulation result of reactance-tuning circuit

Re[Z]
Im[Z]

Impedance (Ω) vs Frequency (GHz)
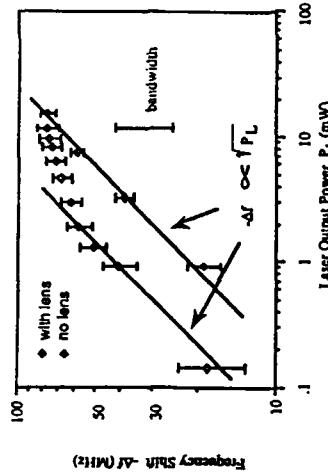
Tunable Active Bandpass Filter Using MESFET as Varactor

- Optical Tuning
  – frequency shift v.s. laser power

(a) constant bias voltages

(b) compensation case
   (varying bias voltages)

Frequency shift – Δf (kHz) vs Laser output power PL (mW)

bandwidth

(a)

(b)
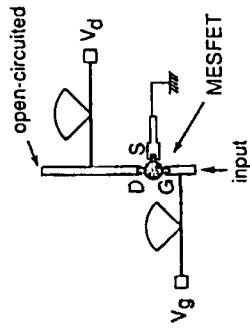
Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor

- MESFET in three-terminal application

S — G D — S

D

S

G
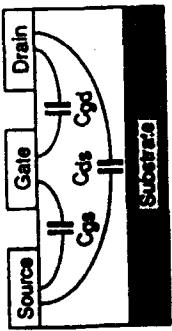
- Three capacitances are involved

Source Gate Drain

Cgs Cds Cgd

Substrate

- electrical-tuning result
- no degradation occurs at Vgr=0
- 170 MHz tuning range
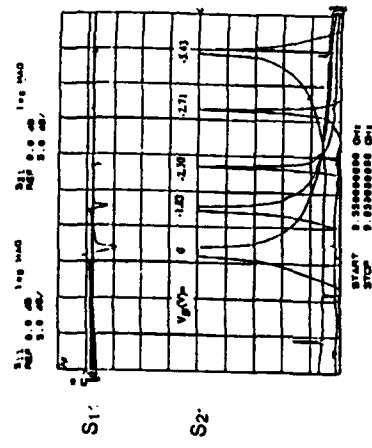
$V_{gr}(V)$ 0   -1.80   -2.30   -2.71   -3.43

START 8.35000000 GHz
STOP 9.05000000 GHz

S_1:
S_2:

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Passive Mode

- electrical-tuning result
- V_{gr} increases
  -> C_{gs} and C_{gd} increase
  -> electrical length increases
  -> frequency decreases

V_{gr} (V)

Frequency (GHz)

... S E Varactor
... Passive Mode

INPUT

OUTPUT

MESFET (varactor)

MESFET (negative resistance)

V_{gv}
V_{dv}
V_{gn}
V_{dn}

G S D   R

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Passive Mode

resonant tank

center line of resonance tank

(a)   (b)

V_{gr} (V)

Frequency (GHz)
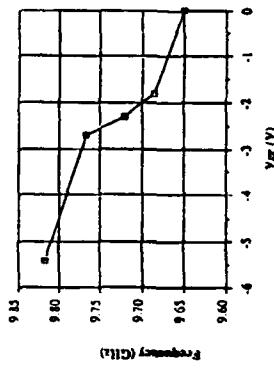
- The position to add the reactance -- (b) is chosen

- Preliminary result from the filter using two-terminal MESFET varactor
- Passband shifts when Vg of the negative resistance MESFET varies from -1.03V to 0V
- Vd of the negative resistance MESFET is kept constant (-2.13V)
- Vgc of the MESFET varactor is kept constant (-2.19V)

$V_g = 0V$
$I_d = 95 mA$

-0.87 -0.46 -0.62 -1.03
50 70 58 45

S_{21} (dB)

0, -10, -20

9.65 9.67 9.69 9.71 9.73 9.75
Frequency (GHz)
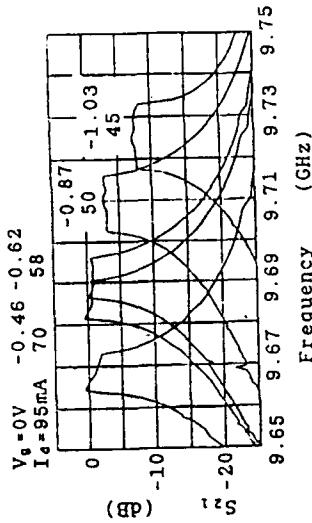
Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor in the
Active Mode

- Preliminary result from the filter using two-terminal MESFET varactor
- Passband shifts when Vd of the negative resistance MESFET varies from 1.61V to 2.34V
- Vg of the negative resistance MESFET is kept constant (-0.46V)
- Vgc of the MESFET varactor is kept constant (-2.20V)

$V_d = 2.34V$ 2.17 1.96 1.84 1.61
$I_d = 72 mA$ 72 71 65 55

S_{21} (dB)

0, -10, -20

9.63 9.65 9.67 9.69 9.71 9.73 9.75
Frequency (GHz)

- optical-tuning result
- no degradation occurs
- 50 MHz tuning range

S_{11}

S_{21}

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Passive Mode

- optical-tuning result
- Laser output power increases
  -> Cgs and Cgd increase
  -> electrical length increases
  -> frequency decreases

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor in the Active Mode

Try to design a tunable active bandpass filter using only one MESFET

• Capacitance variation is larger --> Tuning range is wider

• If Vgs=-1.6V is selected, capacitance variation is optimum

• To keep the insertion loss of active filter at 0dB, the separation of quarter-wavelength coupled line is adjusted from 18 mil to 8 mil

• Then, the gate voltage of the negative resistance needed to keep the insertion loss at 0dB is changed from -0.46V to -1.6V

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor in the Active Mode

• Two MESFETs are replaced by only one MESFET
• The MESFET is biased in the active mode

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor in the Active Mode

Some observations from the preliminary result

1. Tuning range of Vd-tuning is larger than tuning range of Vg tuning.

2. The tuning direction of Vg-tuning can be explained by the depletion-layer capacitance.

   Vg increases --> capacitance increases --> electrical length increases --> frequency decreases

3. The tuning direction of Vd-tuning cannot be explained by the depletion-layer capacitance.

   Vd increases --> capacitance decreases --> electrical length decreases --> frequency increases --> not agree with experimental result ?

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor in the Active Mode

• Active-channel capacitance is important when the MESFET varactor is operated in the active mode.
• Cds is dominant in Vd-tuning

• Measured Cds-Vds relation:

above Vgs=-1.6V the channel is open and the conductance is too large --> cannot measure Cds

1MHz, HP4280A C-V Meter

Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of Vd-tuning
- 120 MHz tuning range

Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of Vd-tuning
- Vd increases
  -> Cgd decreases, but Cds increases and dominates
  -> electrical length increases
  -> frequency decreases
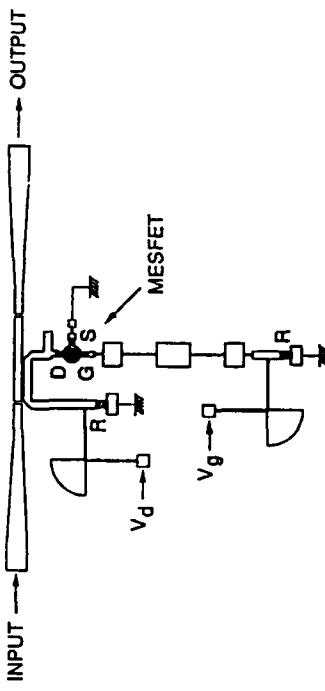
Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of Vg-tuning
- 90 MHz tuning range – smaller than Vd tuning
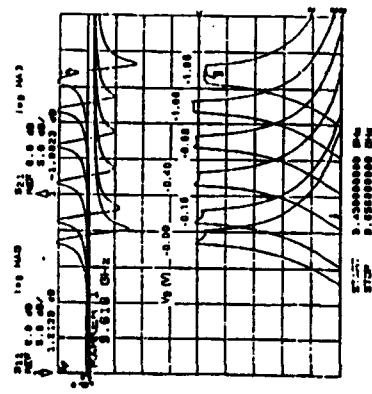
Tunable Active Bandpass Filter Using Three-terminal MESFET Varactor in the Active Mode

- electrical-tuning result of Vg-tuning
- Vg increases
  -> Cgs and Cgd increase
  -> electrical length increases
  -> frequency decreases

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Active Mode

- Optical tuning range is very small (<10MHz)

- The optimum gate bias for optical tuning is above -2V
  e.g. -2.35V for the three-terminal MESFET varactor
  in passive mode

- To keep the insertion loss of the filter at 0dB, a much higher
  gate bias is needed.

- e.g. -1.6V in Vd-tuning
        -1.6V to 0V in Vg tuning

- If the optimum bias condition for the optical tuning is the same as
  the bias condition needed for negative resistance, optical tuning
  becomes practical.
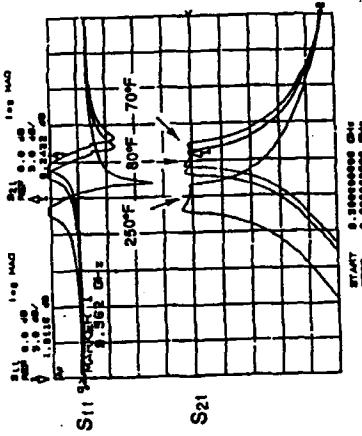
Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor

- Temperature Stability
- 2 MHz shift from 70°F to 80°F

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Active Mode

- electrical-tuning by both Vd and Vg
- 150 MHz tuning range

Tunable Active Bandpass Filter
Using Three-terminal MESFET Varactor
in the Active Mode

- electrical-tuning by both Vd and Vg

Frequency (GHz)	Vd (V)	Vg (V)
0.633	1.65	-1.38
0.610	2.77	-1.38
0.590	4.52	-1.67
0.570	5.64	-1.64
0.550	7.54	-1.64
0.530	8.31	-1.15
0.510	9.73	0
0.490	9.93	0
0.470		

Conclusion

• Coupled Negative Resistance –
  increase the Q value for MMIC application

• Varactor Diode –
  tune the center frequency of the passband

• MESFET Varactor
  1. In addition to the electronic tuning, optical tuning is realized.
  2. Same type of device as the active device generating negative resistance.
  3. Operation in three-terminal mode is investigated.
  4. Show the possibility of using one active device to generate negative resistance and tune the frequency.

RECENT DEVELOPMENTS OF NRD - GUIDE TECHNOLOGY

T. Yoneyama

Tohoku University • Japan
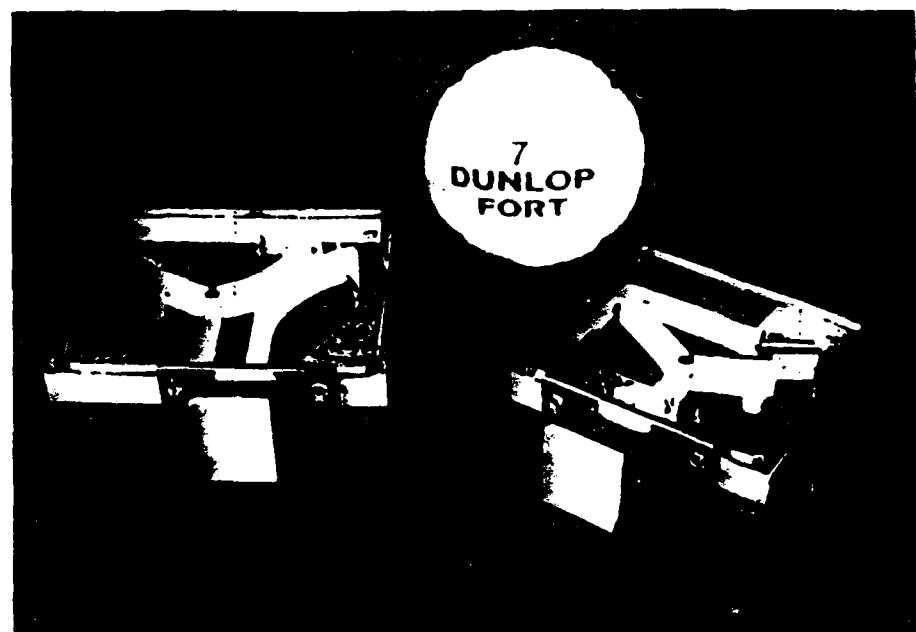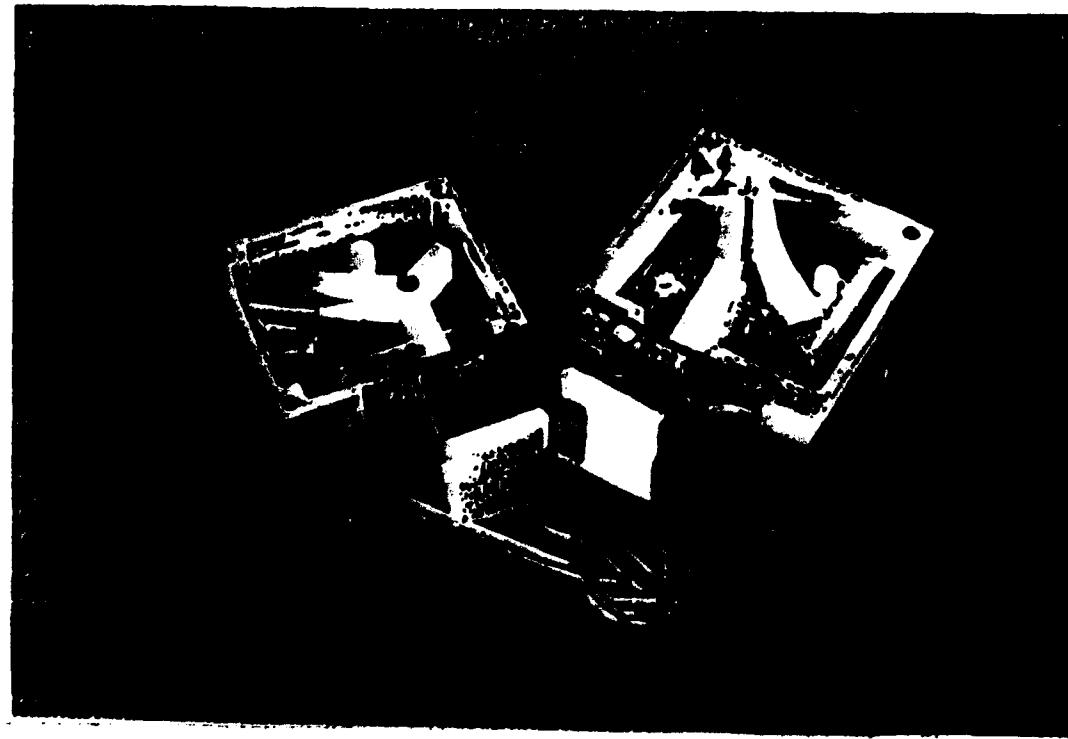
Recent Developement in NRD-Guide
Technology

Tsukasa Yoneyama

Research Institute of Electrical Communication

Tohoku University

Katahira 2-1-1, Aoba-ku

Sendai, 980 Japan

CONTENTS

Metal Plate

Dielectric Strip

LSM₀₁ Mode

LSE₀₁ Mode

—— Electric Field
—— Magnetic Field

(a)   (b)   (c)
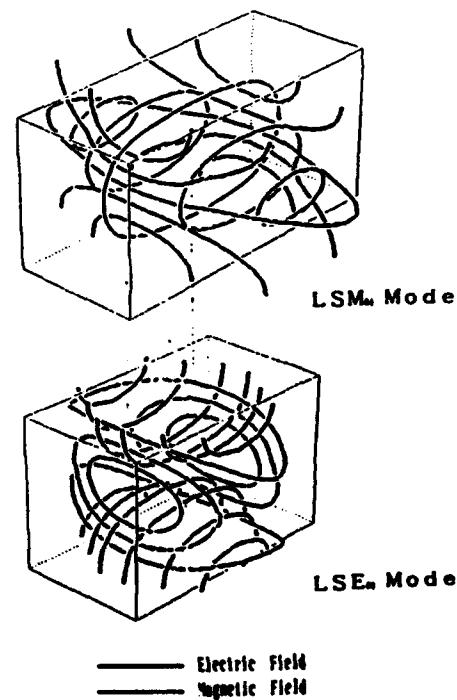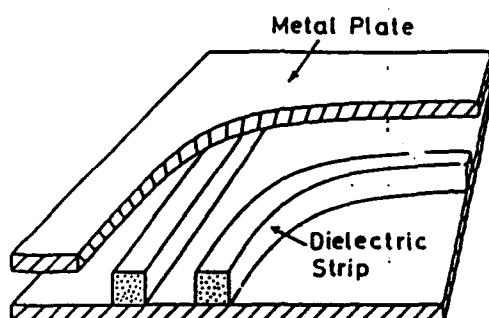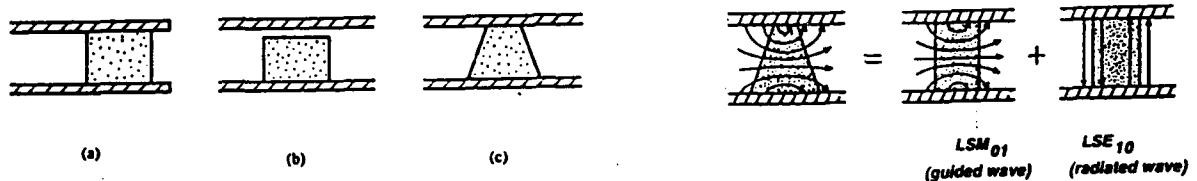
Fig.1 Classical types of leaky wave NRD-guides

LSM_{01} LSE_{10}
(guided wave)   (radiated wave)

Fig.2 Operation principle of leaky wave NRD-guide

5

6
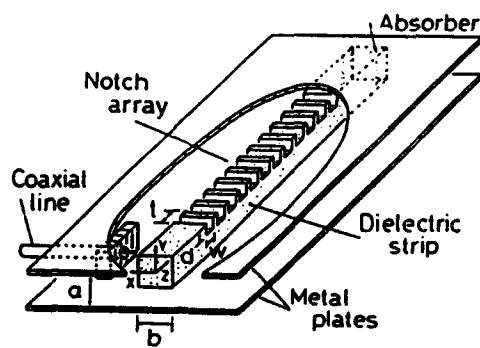
Fig.4 Leaky wave NRD-guide having notches
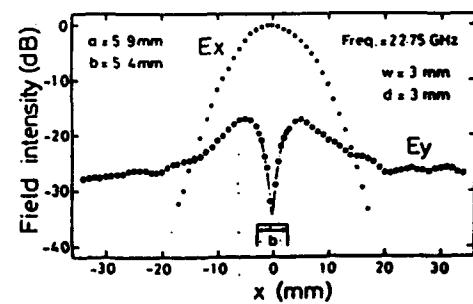
Fig.5 Field distributions in the cross sectional plane
of leaky wave NRD-guide having notches

7

8
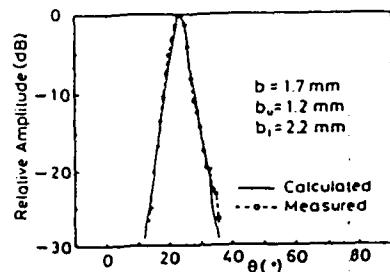
Fig.3 Radiation pattern of leaky wave NRD-guide
having trapezoid dielectric strip

b = 1.7 mm
b_u = 1.2 mm
b_l = 2.2 mm

— Calculated
---•--- Measured

(a) Upper part

(b) Lower part
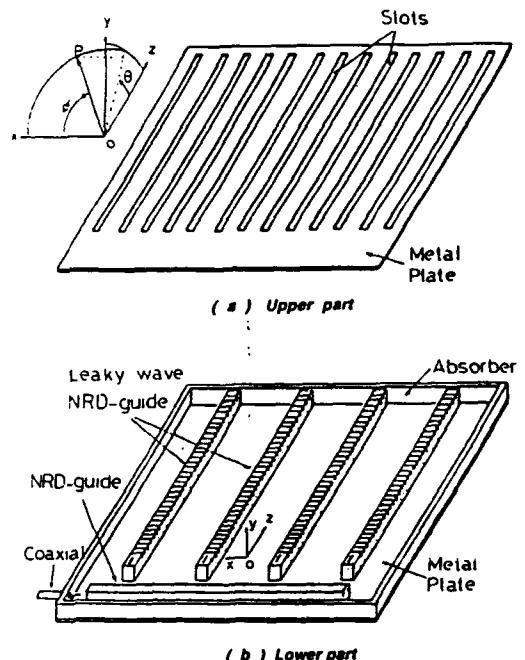
Storucture of leaky wave NRD-guide fed planar anntena

z = 130mm
z = 180mm
=22·75GHz
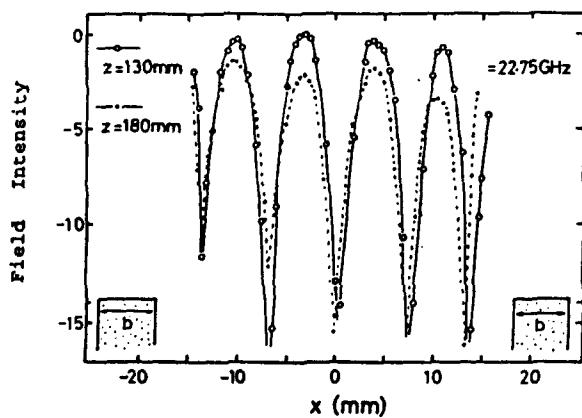
Distribution of vertical component of E-field in the planar
antenna

12

13

(a) θ方向特性

(b) φ方向特性
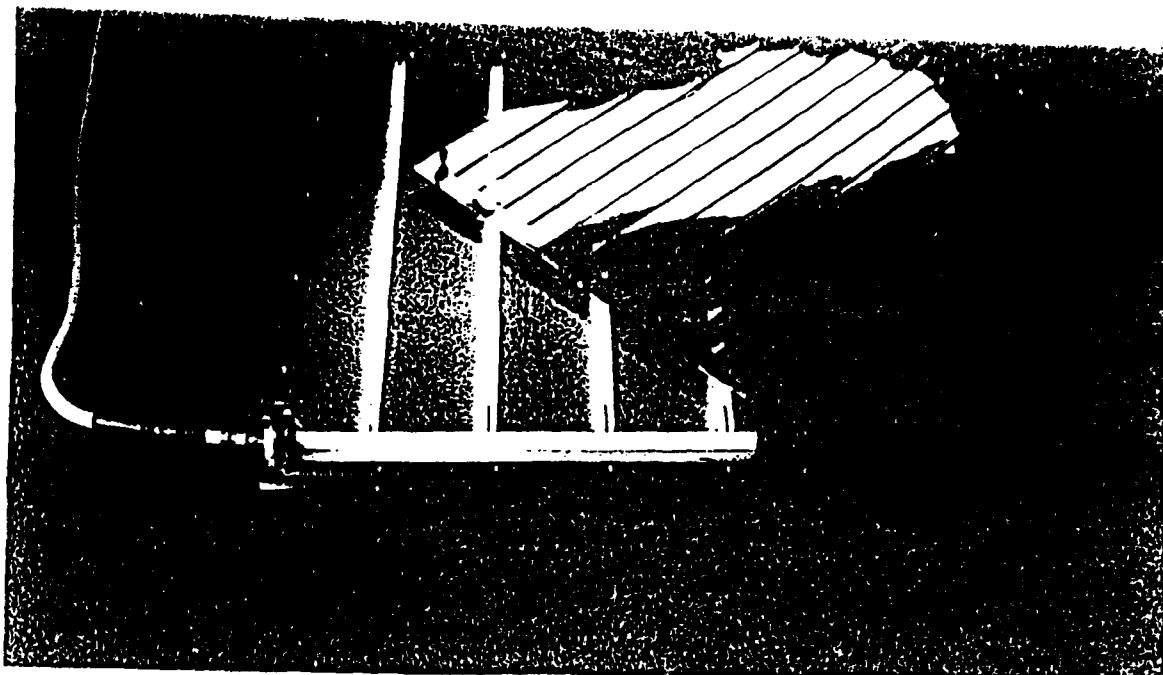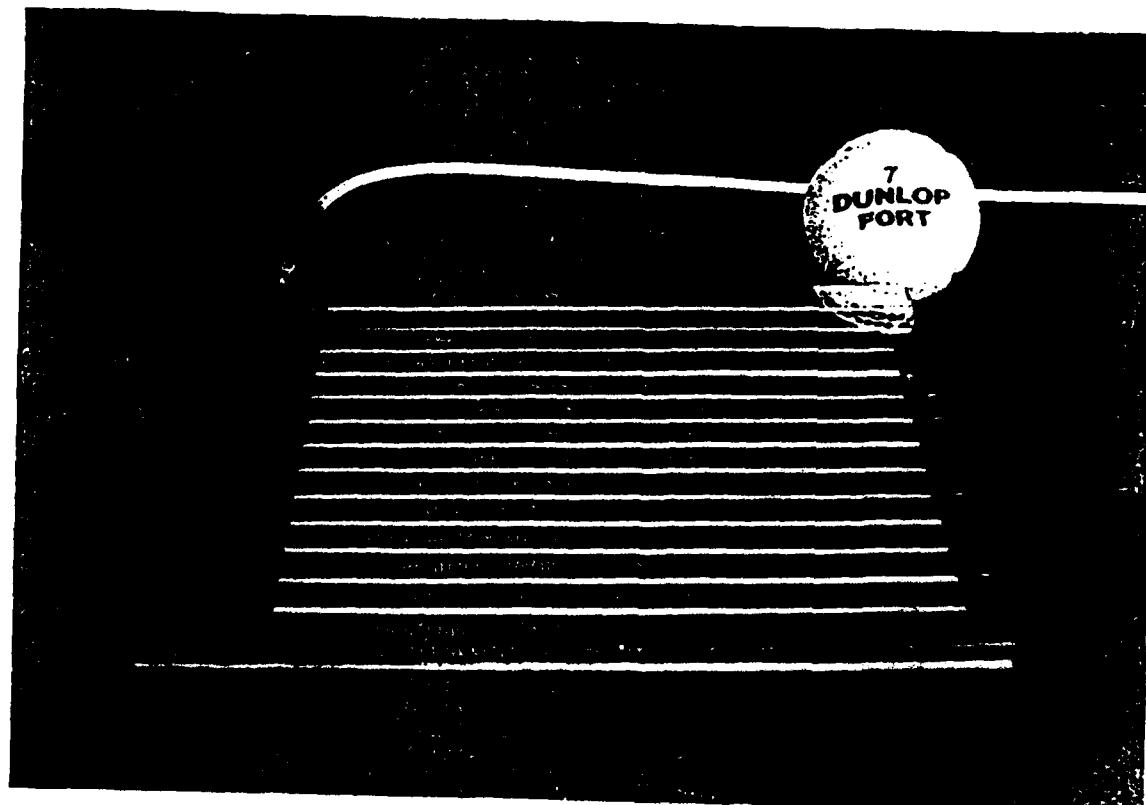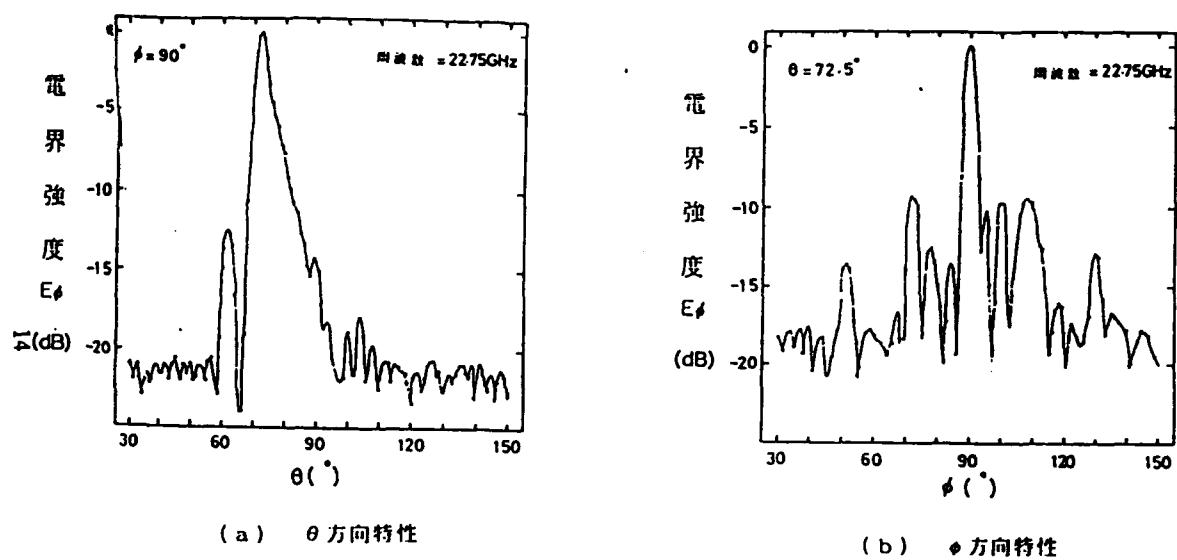
図7 漏れ波NRDガイド給電平面アンテナの放射指向性
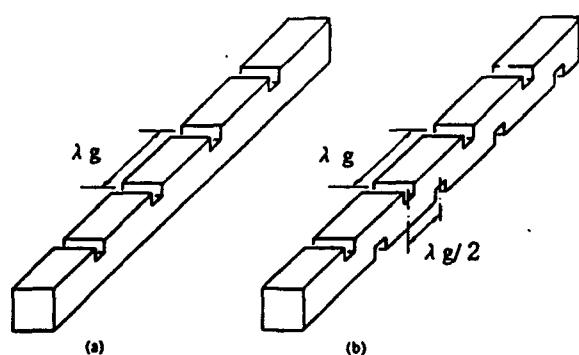
Fig.6 Structure of broadside leaky wave NRD-guide

15
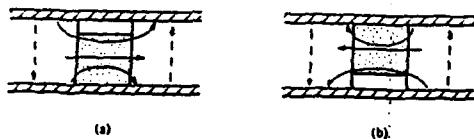
Fig.7 Field distributions of broadside leaky wave NRD- guide
at notches separated by half a wavelength
(solid line : guided wave, dotted line : radiated wave)

a = 5.9 mm
b = 5.4 mm
w = 2.0 mm
d = 2.0 mm
t = 23.0 mm   Freq. = 23.9 GHz
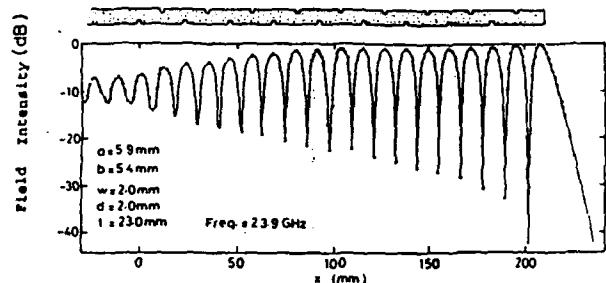
Standing wave pattern along broadside leaky wave NRD-
guide

16

17
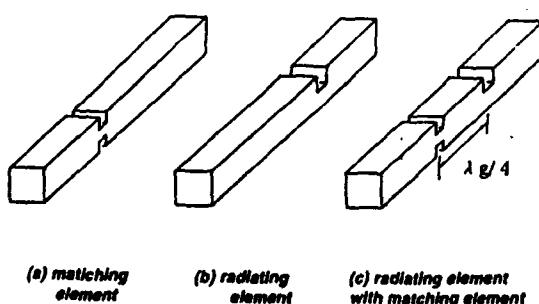
(a) matching   (b) radiating   (c) radiating element
    element   element   with matching element

Fig.8 Matching element and radiating element
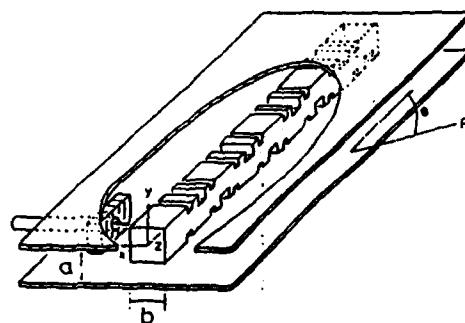
Fig.9 Structure of broadside leaky wave NRD-guide

18

19

Radiating notch
w = 2.5 mm
d = 2.5 "
t = 24.0 "

Matching notch
w = 2.0 mm
d = 2.0 "
t = 24.0 "

Freq = 24 GHz
a = 5.9 mm
b = 5.4 mm

Antenna length
= 180 mm

Relative Amplitude (dB)

θ (°)

Fig.10 Radiation pattern of broadside leaky wave NRD-guide

Slots

Leaky wave
NRD-guide

Coaxial
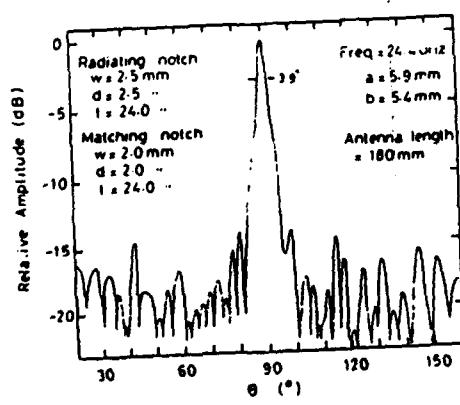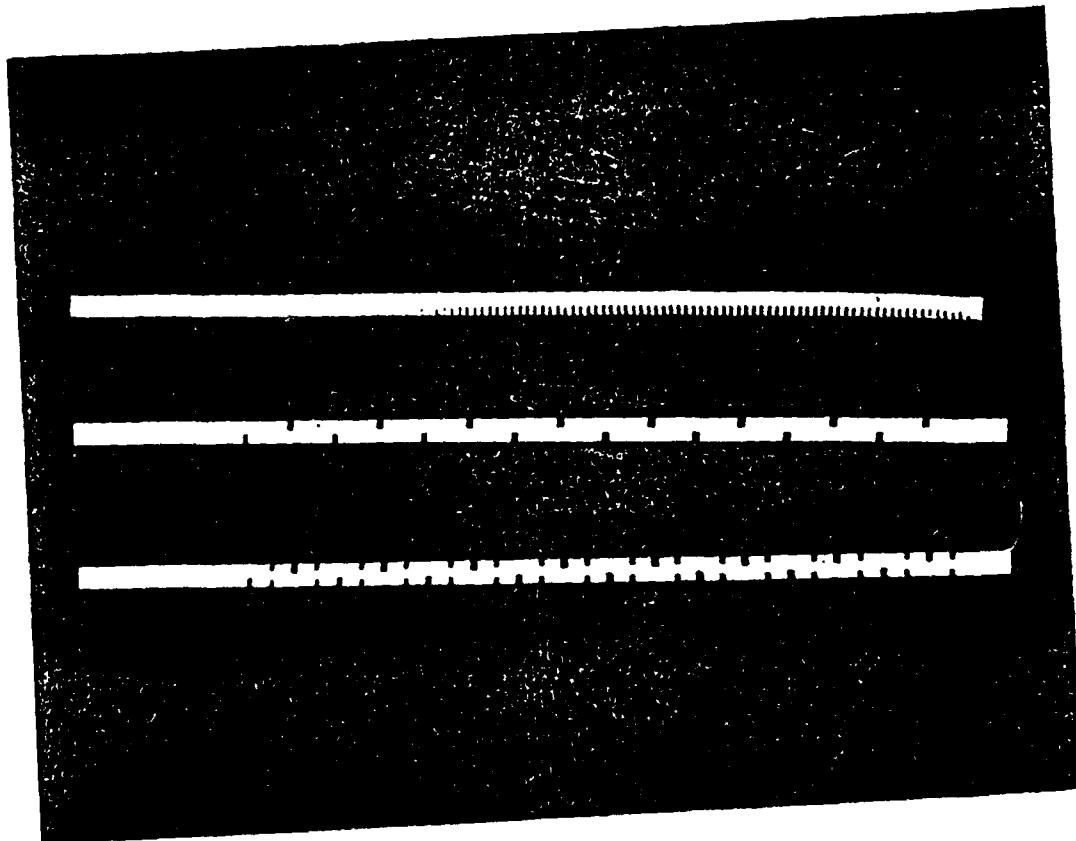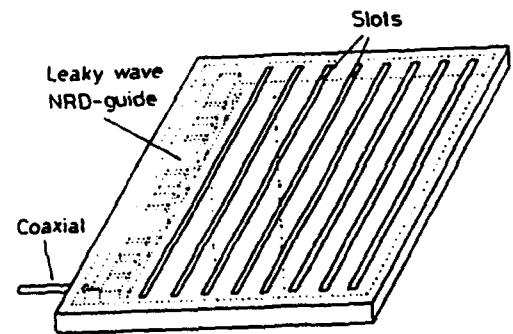
Fig.11 Planar antenna fed by leaky wave NRD-guide

23
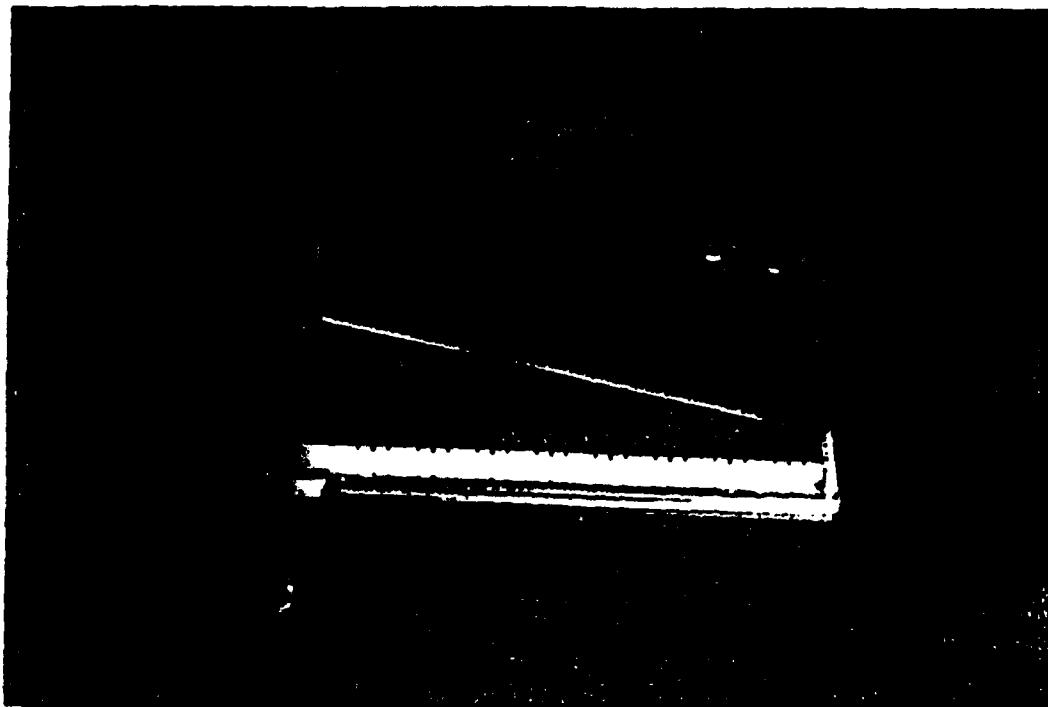
24
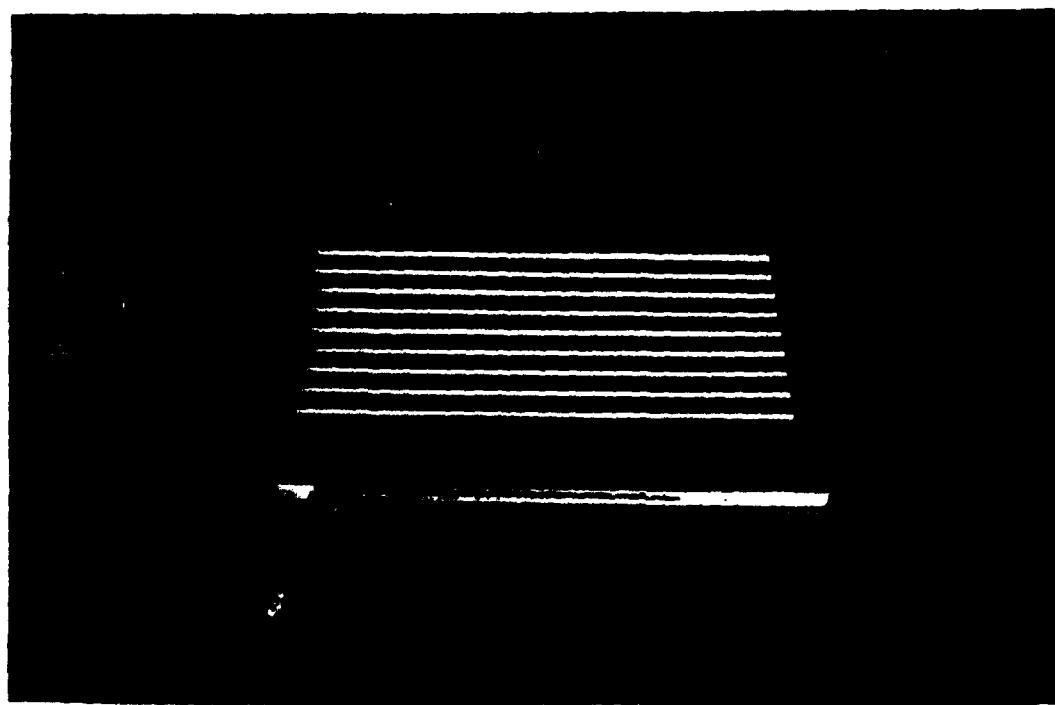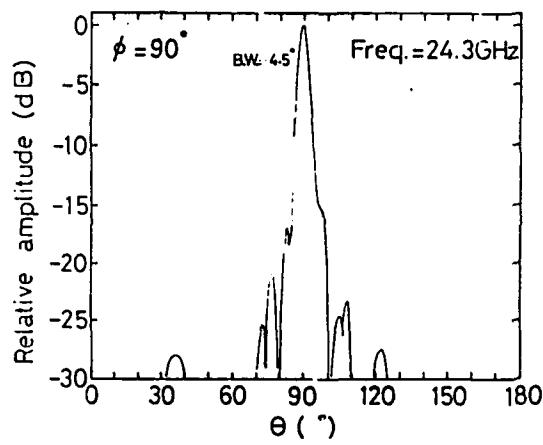
Fig.12-(a) Radiation pattern of broadside leaky wave NRD-
guide fed planar anntena (230x150x7 in mm)
(in the vertical plane parallel to the slot)
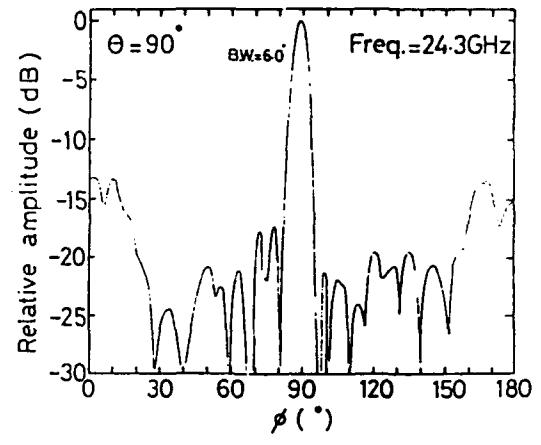
Fig.12-(b) Radiation pattern of broadside leaky wave NRD-
guide fed planar anntena (230x150x7 in mm)
(in the vertical plane perpendicular to the slot)

25

26

27
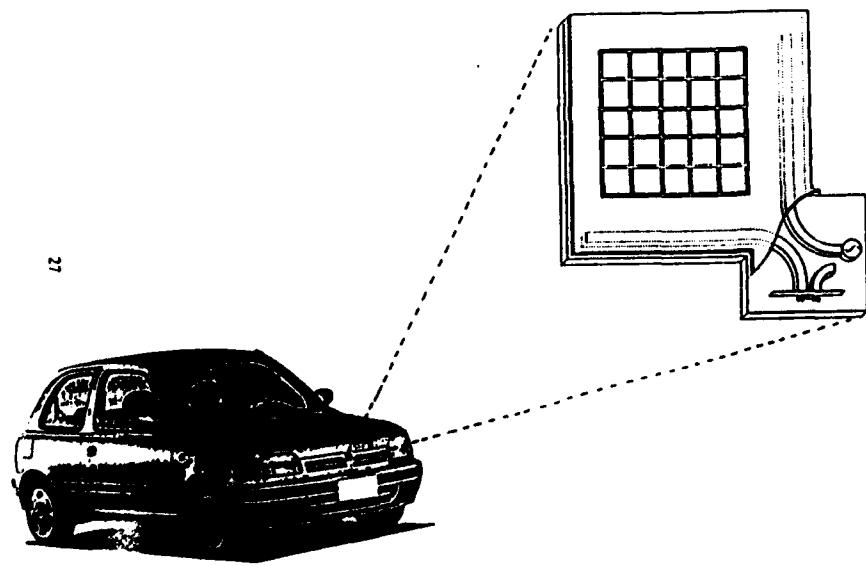
ADVANCED DESIGN TECHNIQUES

FOR LINEAR AND NON LINEAR MMIC'S INTO THE

MILLIMETER - WAVE REGION

R. H. Jansen

Jansen Microwave • Germany

ADVANCED DESIGN TECHNIQUES FOR LINEAR AND NONLINEAR MMICS

INTO THE MM-WAVE REGION

Rolf H. Jansen, Jansen Microwave, Ratingen, Germany

SUMMARY

In the past few years, the field of high-speed and microwave monolithic integrated circuits (MMICs) has seen an increasing range of nonmilitary applications. Particularly large growth rates and competitive development prevails in the emerging markets of mobile and satellite communications as well as of automotive electronic equipment up to high mm-wave frequencies. Commercial pressure is now driving the demand for faster development and lower cost manufacturing of GaAs MESFET-based and HEMT-based MMICs. From a design point of view, this is reflected in the following trends:

- The requirement for high packing density circuits, i.e. MMICs with minimized chip size, and for multifunction circuits integrating several functional blocks on a single chip while maintaining reasonably small size and thus acceptable yield. This is paralleled by the creation of miniaturized lumped components working well into the mm-wave region.

- The tendency towards lower noise figures and higher frequencies through the gradual transition from MESFET to HEMT technologies and in view of mm-wave applications like collision-avoidance radars. Several MMIC foundry services are now in the stage of moving to higher frequency limits, i.e. from about 20 GHz presently to 40...60 GHz in the near future.

- The demand for better CAD capable of addressing the problems associated with high packing density design and with high mm-wave frequencies in process-related form. The goal here is the reduction of development time and the full exploitation of the degrees of freedom offered by the MMIC medium for linear and nonlinear circuit design.

Under contract with GEC-Marconi Materials Technology, UK, formerly Plessey Research Caswell, the author and his group have worked over a number of years on such as the above outlined issues. Progress achieved is visualized by representative examples. A brief overview is given over advances in CAD techniques directly related to the MMIC process technology. This includes third generation look-up table concepts for single and dual metallization with high speed and for high geometrical complexity. A unified approach

has been made available here for microstrip type MMICs and for coplanar waveguide type (CPW) circuits. For specific, repeatedly used structures like via holes, approximate field-theory based models have been developed for fast interactive CAD. Further, linear and nonlinear distributed FET/HEMT modelling in terms of device layout geometry is described, including automated parameter extraction. CAD based on this approach has near interactive speed and has primarily been developed for frequency domain design involving larger devices and for analogue MMICs operating at high mm-wave frequencies.

REFERENCES related to this workshop contribution

/1/ Hente, D. and Jansen, R.H.: A frequency-domain continuation method for the analysis and stability investigation of nonlinear microwave circuits, IEE Proc., Vol. 133, Pt.H., Oct. 1986, 351-362.

/2/ Jansen, R.H.: Recent advances in the full-wave analysis of transmission lines for application in MIC and MMIC design, Invited Paper, Proc. SBMO Intern. Microwave Symposium, Rio de Janeiro, July 1987, 467-475.

/3/ Jansen, R.H. and Wertgen, W.: A 3D field-theoretical simulation tool for the CAD of mm-wave MMICs, Alta Frequenza, Italy, vol. LVII, 1987, 203-216.

/4/ Jansen, R.H. et al: A comprehensive CAD approach to the design of monolithic microwave integrated circuits, Invited Paper, IEEE Trans., MTT-36, (1988), 208-219.

/5/ Wiemer, L. and Jansen, R.H.: Reciprocity-related definition of strip characteristic impedances for multiconductor hybrid-mode transmission lines. Microwave and Optical Technology Letters, Wiley Interscience, vol. 1, (1988), 22-25.

/6/ Jansen, R.H., Eddison, I.G., Arnold, R.G.: Recent developments in the CAD of high packing density MMICs, Microwave and Optronics Conf (MIOP), 1989, Sindelfingen, Germany, paper 1/3.

/7/ Jansen, R.H. and Wiemer, L.: Full-wave theory based development of mm-wave circuit models for microstrip open end, gap, step, bend and tee, IEEE MTT-S Digest, Long Beach, USA, 1989, 779-782.

/8/ Jansen, R.H.: Full-wave analysis and modeling for CAD of mm-wave MMICs, Proc. Intern. Workshop on Millimeter Waves, April 1989, Perugia, Italy, Paper No. 2, also in Alta Frequenza, vol. LVIII, 1989, 551-558.

2

/9/ Wertgen, W. and Jansen, R.H.: Efficient direct and iterative electrodynamic analysis of geometrically complex MIC and MMIC structures, Intern. Journal of Numerical Modelling, Wiley, vol. 2, 1989, 153-186.

/10/ Jansen, R.H.: Advanced CAD/CAM software for process-related MMIC design, Proc. ESA Workshop on MMICs for Space Applic., Noordwijk, March 1990, paper 7/2.

/11/ Jansen, R.H.: LINMIC+, A European familiy of CAD tools for MIC and MMIC design, Microwave Engineering Europe, March/April 1991, 21-24.

/12/ Brömme, R. and Jansen, R.H.: Systematic investigation of coplanar waveguide MIC/MMIC structures using a unified strip/slot 3D electromagnetic simulator, IEEE MTT-S Digest, Boston, USA, 1991, 1081-1084.

/13/ Jansen, R.H. and Sauer, J.: High-speed 3D electromagnetic simulation for MIC/MMIC CAD using the spectral operator expansion (SOE) technique, IEEE MTT-S Digest, Boston, USA, 1991, 1087-1091.

/14/ Jansen, R.H. and Pogatzki, P.: Nonlinear distributed modelling of multifinger FETs/HEMTs in terms of layout geometry and process data, Proc. 21st Europ. Microwave Conf., Stuttgart, Germany, Sept. 1991, 609-614.

/15/ Jansen, R.H.: CAD oriented spectral domain modelling of MIC/MMIC structures, Intern. Workshop IEEE MTT/AP Germany Digest, Stuttgart, Sept. 1991, 19-31.

/16/ Jansen, R.H.: Novel, process-related design techniques for linear and nonlinear MMICs, 1991 Asia-Pacific Microwave Technology and Education Workshop, Taiwan, Sept. 1991, paper TA3, 65 pages.

/17/ Jansen, R.H.: A full-wave electromagnetic model of cylindrical and conical via hole grounds for use in interactive MIC/MMIC design, IEEE MTT-S Int. Microwave Symposium, Albuquerque, USA, 1992, accepted for publication.

CAD OF MILLIMETER - WAVE PASSIVE COMPONENTS

IN SUSPENDED MICROSTRIP TECHNOLOGY

V. Fouad Hanna

CNET • France

P L A N

* INTRODUCTION.

* EVALUATION OF EIGENMODES IN SUSPENDED STRIPLINE (SSL) TECHNOLOGY.

* CARACTERISATION OF UNIAXIAL SSL DISCONTINUITIES.

* REALISATION OF PASSIVE MICROVE COMPONENTS IN SSL TECHNOLOGY : Directional Coupler-Resonator-Filter.

* CARACTERISATION OF MULTIAXIAL SSL DISCONTINUITIES : Bend , Stub ,Bent-Stub.

SPECTRAL DOMAIN APPROACH
for a SSL

$$E_z(x,y,z,t) = \sum_{\alpha_n} E_z(\alpha_n,y) \cdot e^{j\alpha_x x} \cdot e^{j(\omega t - \beta z)} \quad (1)$$

where

$$\alpha_n = n\frac{\pi}{B}$$
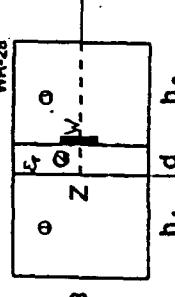
$$\begin{bmatrix} E_x(\alpha_n,d) \\ E_z(\alpha_n,d) \end{bmatrix} = \begin{bmatrix} G_{11}(\alpha_n,\beta,\omega) & G_{12}(\alpha_n,\beta,\omega) \\ G_{21}(\alpha_n,\beta,\omega) & G_{22}(\alpha_n,\beta,\omega) \end{bmatrix} \begin{bmatrix} J_x(\alpha_n,d) \\ J_z(\alpha_n,d) \end{bmatrix} \quad (2)$$

WHY?

	Stripline (microstrip)	Suspended Stripline (suspended microstrip)
Attenuation losses	High	75% Less than microstrip
Dispersion	Moderate	Little
C A D Facilities	Rich	Poor

WHY?

Parallel-coupled Band-pass filter

Coupled-gap Band-pass filter

An Asymmetric Suspended Stripline

Fig.3

CAD OF MILLIMETER-WAVE PASSIVE COMPONENTS IN SUSPENDED MICROSTRIP TECHNOLOGY

Victor FOUAD HANNA*, Joseph ACHKAR*, Odile PICON* Jacques CITERNE**

*DIVISION
SYSTEMES DE TELECOMMUNICATIONS PAR SATELLITE
CENTRE NATIONAL D'ETUDES DES TELECOMMUNICATIONS

PARIS, FRANCE

**INSA Rennes, France

SUSPENDED AND INVERTED MICROSTRIP LINES

- Reduced attenuation (detached ground plane)

- Millimeter applications ──► Shielding

- Field configurations TEM (closer than microstrip)

- Circuit layout on both sides of the substrate.

Fig. 1

Fig 2 - a

Fig 2-b
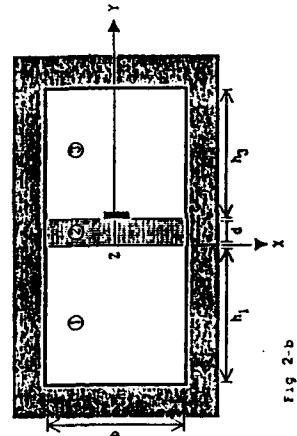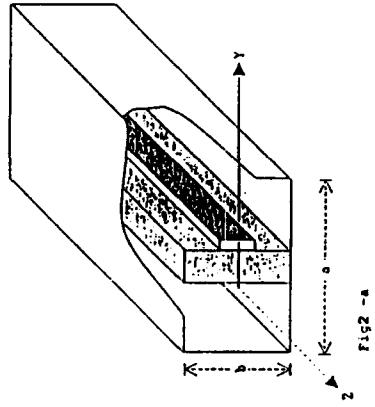
Longitudinal (1-a) and cross-sectional (1-b) view of suspended stripline.

SDA

[15]

Z_{oo}

Z_{oe}

W/H

Z (Ohms)
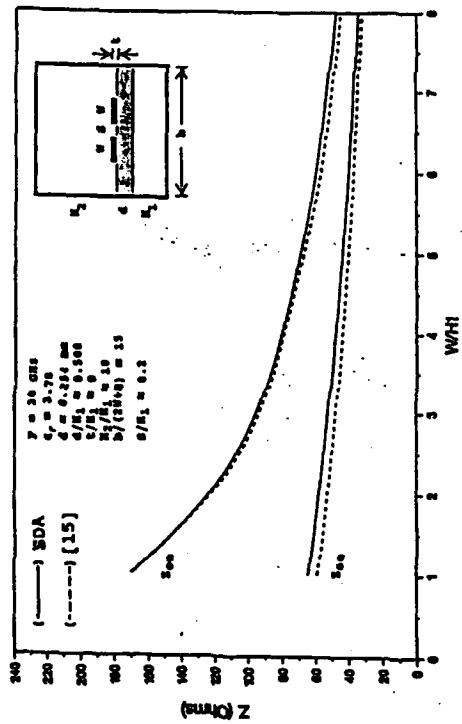
Fig 6 Variation of the characteristic impedance for the even and odd modes as a function of W/H for coupled SSL's: (—) our analysis, (---) reference [15].

SDA

[14]

W (mm)

Z (Ohms)
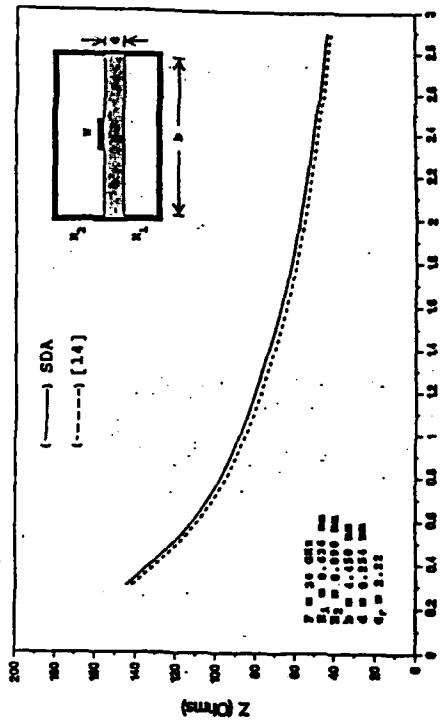
Fig 7 Variation of the characteristic impedance as a function of W for an SSL: (—) our analysis, (---) reference [14].

F (GHz)

β, $|\beta|$ (rd/m)
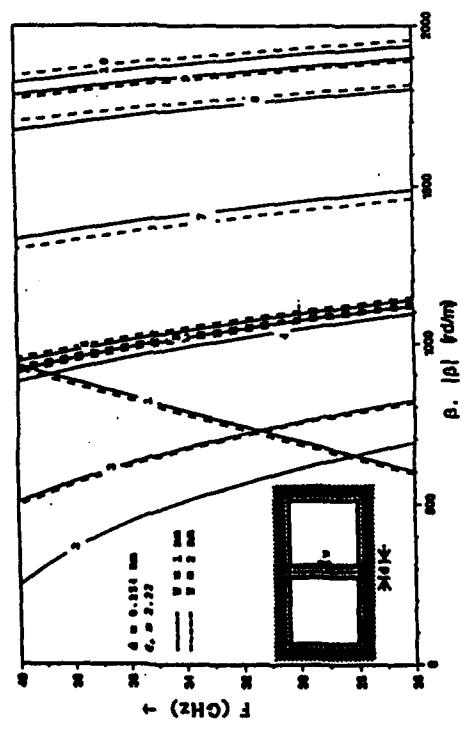
Fig 4 Dispersion characteristics of symmetric SSL's ($\epsilon_r = 2.22$).

F (GHz)

β, $-j\beta$ (rd/m) →
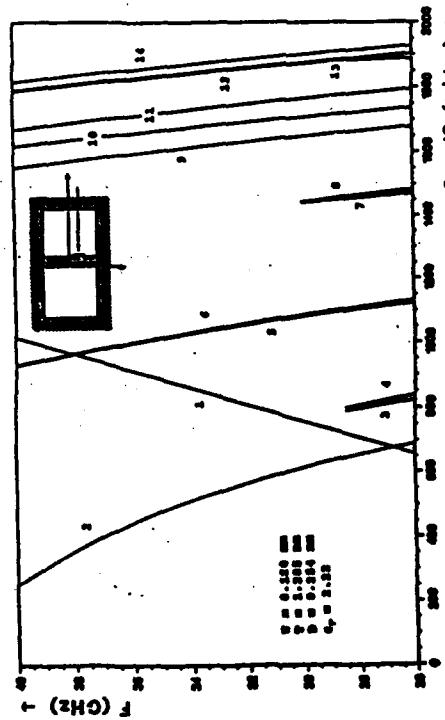
Fig 5 Dispersion characteristics of asymmetric SSL ($\epsilon_r = 2.22$).

	$F(x)$	$J(a_z)$
J_{z1}	$1 \quad |x| \le \frac{w}{2}$ $0 \quad |x| > \frac{w}{2}$	$\frac{2}{a_z \cdot b} \cdot \sin a_z \cdot \frac{w}{2}$
J_{z2}	$\frac{1}{\sqrt{1 - \left(\frac{2x}{w}\right)^2}}$ 0	$\frac{\pi w}{2b} \cdot J_0\left(a_z \cdot \frac{w}{2}\right)$
J_{x1}	$\frac{2x}{w}\sqrt{1 - \left(\frac{2x}{w}\right)^2}$ 0	$-j \frac{\pi}{a_z \cdot b} \cdot J_1\left(a_z \cdot \frac{w}{2}\right)$
J_{x2}	$\frac{\sin\left(4\pi\frac{x}{w}\right)}{\sqrt{1 - \left(\frac{2x}{w}\right)^2}}$ 0	$-j\frac{\pi w}{b}\left[J_1\left(2\pi - a_z \cdot \frac{w}{2}\right) - J_1\left(2\pi - a_z \cdot \frac{w}{2}\right)\right]$

$a_z = (2n-1)\pi/b$

T-1.1

Method of GALERKIN

$$J_z(x,d) = \sum_{n=1}^{\infty} a_z \cdot J_{z,n}(x,d) \quad \Rightarrow \quad J_z(a_z,d) = \sum_{n=1}^{\infty} a_z \cdot J_{z,n}(a_z,d) \qquad (3)$$

$$J_x(x,d) = \sum_{n=1}^{\infty} b_z \cdot J_{x,n}(x,d) \quad \Rightarrow \quad J_x(a_z,d) = \sum_{n=1}^{\infty} b_z \cdot J_{x,n}(a_z,d)$$

Relationship of Complementarity

$$\int_{-\infty}^{\infty} E_z(x,d) \cdot J_z(x,d) \cdot dx = 0 \quad \Rightarrow \quad \text{(R+S) homogeneous}$$
$$\int_{-\infty}^{\infty} E_z(x,d) \cdot J_x(x,d) \cdot dx = 0 \quad \Rightarrow \quad \text{linear equations}$$
(4)

$$\text{Det}(\beta, f, \text{geometry}) = 0$$

$$\beta^2 > 0 \quad \leftarrow \qquad \rightarrow \quad \beta^2 < 0 \qquad (5)$$

propagating modes $\qquad$ evanescant modes

Discontinuity Boundary Conditions

$$\bar{E}_T^{I} = \sum_{p=1}^{P}(A_p^{I} + B_p^{I}) \cdot \bar{e}_{p,T}^{I}$$

$$\bar{H}_T^{I} = \sum_{p=1}^{P}(A_p^{I} - B_p^{I}) \cdot \bar{h}_{p,T}^{I} \qquad (6)$$

$$\bar{E}_T^{II} = \sum_{q=1}^{Q}(A_q^{II} + B_q^{II}) \cdot \bar{e}_{q,T}^{II}$$

$$\bar{H}_T^{II} = \sum_{q=1}^{Q}(B_q^{II} - A_q^{II}) \cdot \bar{h}_{q,T}^{II} \qquad (7)$$

Discontinuity Boundary Conditions

on S_I: $\quad \bar{E}_T^{I} = 0$

$$\bar{H}_T^{II} + \bar{J}_T^{I} \times \bar{u} = 0 \qquad (8)$$

on $S_{II} - S_I$:

$$\bar{E}_T^{I} = \bar{E}_T^{II}$$
$$\bar{H}_T^{I} = \bar{H}_T^{II} \qquad (9)$$

$\Rightarrow$ Set of 2(P+Q) linear equations $\Rightarrow$ 2(P+Q) Modal Amplitudes

Generalised Discontinuity Scattering Matrix

$$[B] = [S] \cdot [A] \qquad (10)$$

$$[B] = \begin{bmatrix} [B_P^{I}] \\ [B_Q^{II}] \end{bmatrix} \qquad [A] = \begin{bmatrix} [A_P^{I}] \\ [A_Q^{II}] \end{bmatrix}$$

$$[S] = \begin{bmatrix} [S_{11}] & [S_{12}] \\ [S_{21}] & [S_{22}] \end{bmatrix}$$

a step discontinuity in an SSL
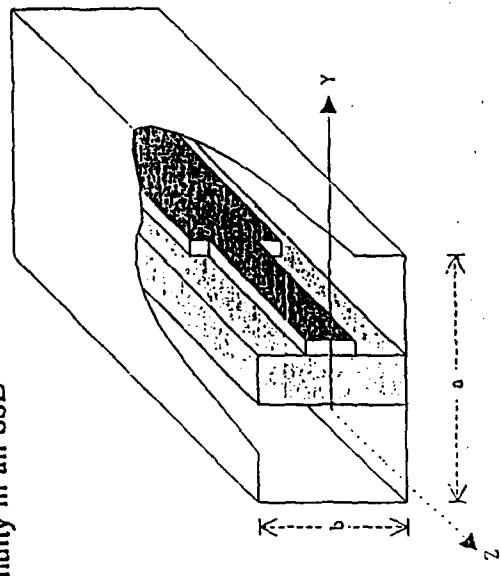
Fig 8 Variation of the x component of the electric field as a function of y (mm) in an SSL
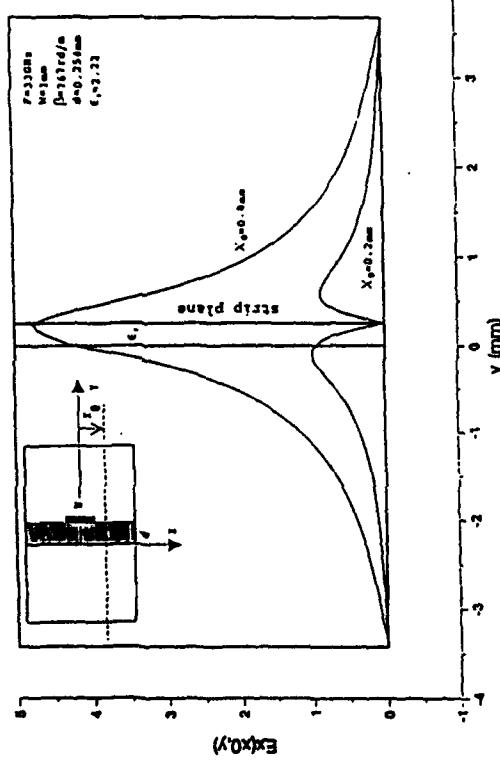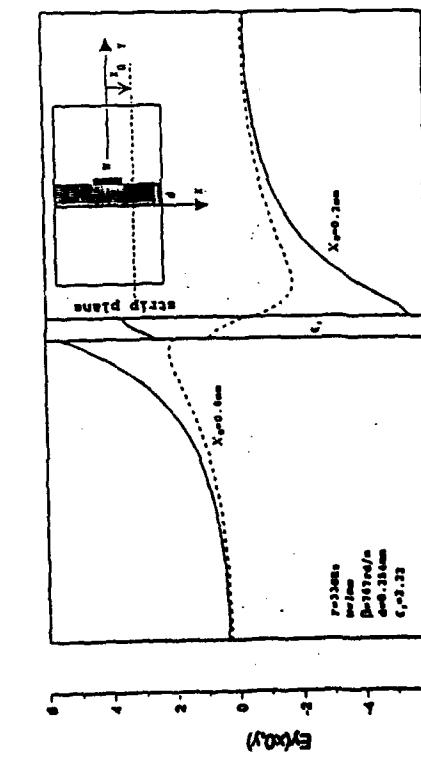
Fig 9 Variation of the y component of the electric field as a function of y (mm) in an SSL

Fig 10

Fig 15 Variation of the equivalent capacitances as a function of frequency in the equivalent circuit for an SSL Gap discontinuity.
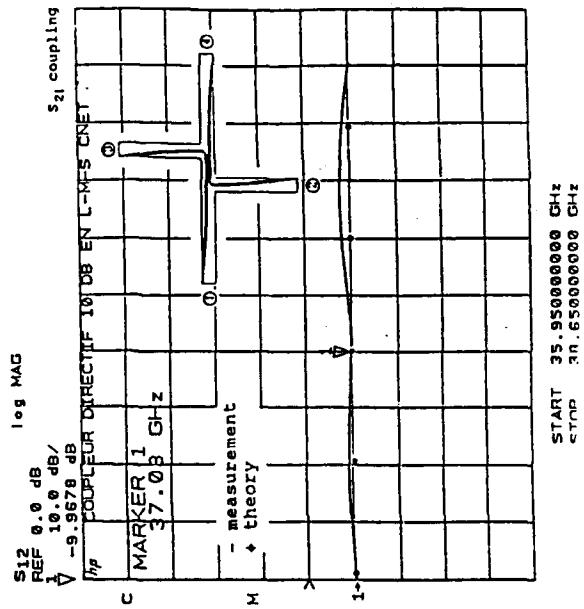
Fig 16 Theoretical and experimental variation of the coupling coefficient of the directional coupler as a function of frequency in the Ka band.

Fig 13 Variation of the equivalent end-effect length as a function of frequency in the Ka band for an open end asymmetric SSL
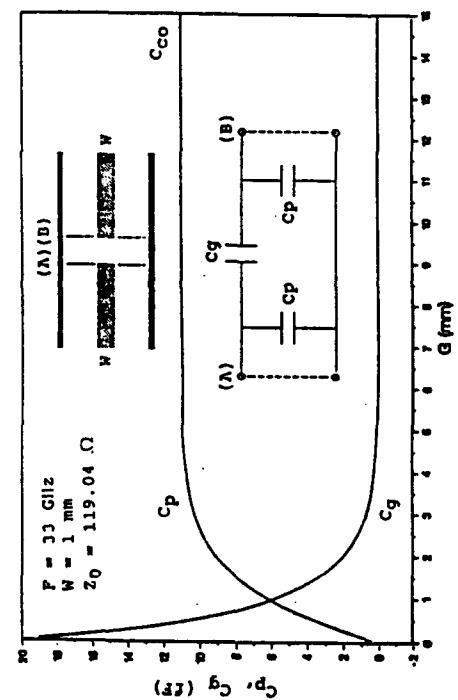
Fig 14 Variation of the equivalent capacitances as a function of the Gap's width in the equivalent circuit for an SSL Gap discontinuity.

Role of the Spectral Domain Approach

se constants.

amplitudes of eigenmode field components.

pling coefficients $\Big\}$ directly in the spectral domain

er flow

$$P_{p,q} = \frac{1}{2}\oint_S (\vec{E}^I \wedge \vec{H}^{II}) \cdot d\vec{S} \quad (11)$$

I II
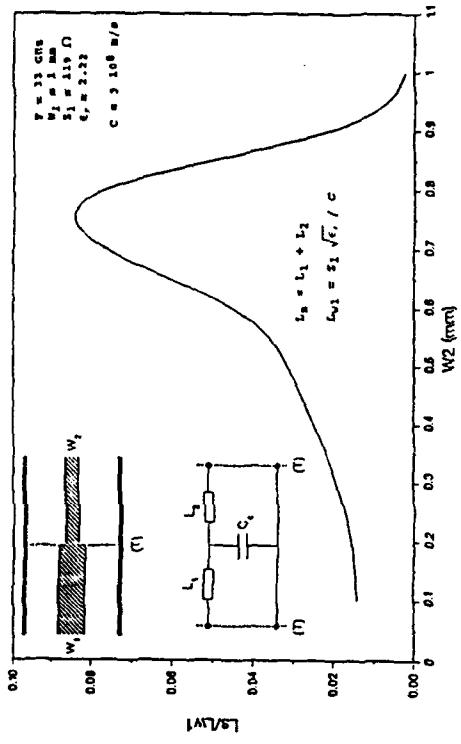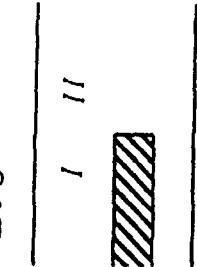
Fig 11 Variation of the equivalent normalized *inductance as a function of the strip width in the equivalent circuit for an SSL step discontinuity.*
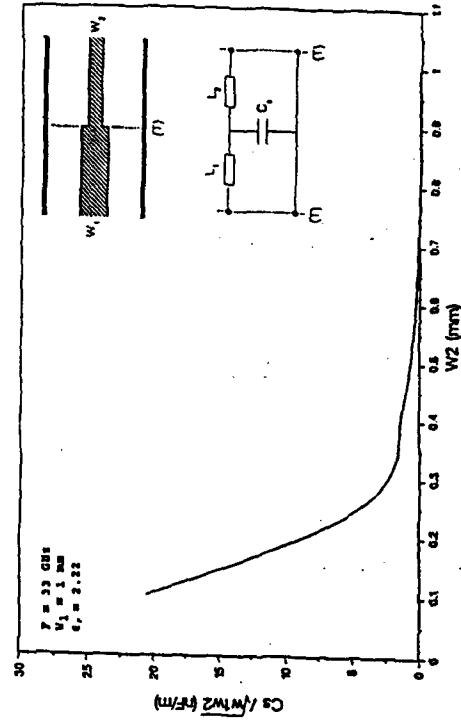
Fig 12 Variation of the equivalent normalized capacitance *as a function of the strip width in the equivalent circuit for an SSL step discontinuity.*

INTEGRAL EQUATIONS ⟹ MATRIX EQUATION

* CONDUCTOR RECTANGULAR CELLS

* CURRENT ⟶ OVERLAPPING

$$\vec{J}_s(\vec{r}) = \frac{1}{W}\sum_{N=1}^{N-1} I_n \vec{T}(\vec{r}).\vec{e}_x \qquad (7)$$

$$I_s = \frac{-1}{j\omega}|\vec{T}_1, \vec{J}_s|$$

Test function

Current cell

Charge cell
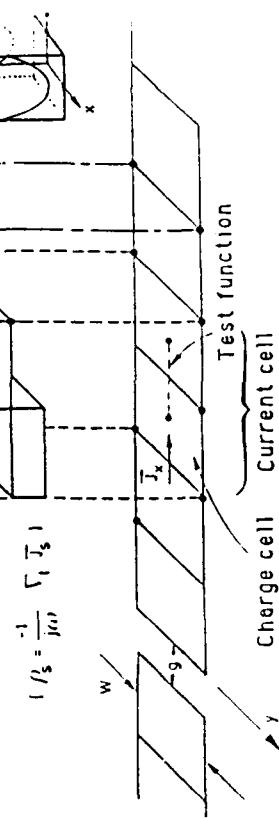
Fig 22

SIMULATION OF A VIRTUAL MATCHED LOAD

The obtained new matrix equation, for the loaded structure, can be written in the form :

$$[[Z] + [\Delta Z]][I] = [V]$$

where [Z] is the generalized impedance matrix of the unloaded structure. [I] is the unknown current vector, [V] is the feeding voltage vector and [ΔZ] is a square diagonal impedance matrix.
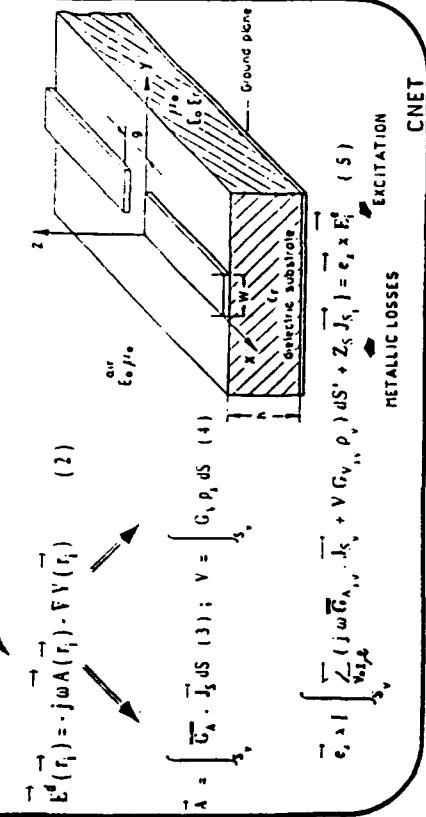
- INTEGRAL EQUATIONS TECHNIQUES.

- METHOD OF MOMENT.

- MATCHED LOAD SIMULATION.

Fig 2 (a) Top view. (b) equivalent circuit (c) current distribution

(d) A normalized current load mounting of the N last cellules

ANALYSIS

$$\vec{e}_z \times [\vec{E}^d(\vec{r}_1) + \vec{E}^e(\vec{r}_1)] = \vec{e}_z \times [Z_s \vec{J}_s(\vec{r}_1)] \qquad (1)$$

$$\vec{E}^d(\vec{r}_1) = -j\omega\vec{A}(\vec{r}_1) - \nabla V(\vec{r}_1) \qquad (2)$$

$$\vec{A} = \int_{S_v} \overleftrightarrow{G}_A \cdot \vec{J}_s \, dS \quad (3) \; ; \quad V = \int_{S_v} G_v \rho_v \, dS \quad (4)$$

$$\vec{e}_z \times \int_{S_v} \left[\sum_{v=1}^{N} (j\omega\overleftrightarrow{G}_{A_{1v}} \cdot \vec{J}_{s_v} + \nabla G_{v_{1v}} \rho_v) \, dS' + Z_s \vec{J}_{s_1} \right] = \vec{e}_z \times \vec{E}^e_1 \qquad (5)$$

METALLIC LOSSES

EXCITATION

CNET

$|S_{11}|$

$|S_{21}|$

Fig 19 Variation of S11 (dB) and S21 (dB) as a function of the resonator length at F=28.5GHz.

Fig 20 Theory, simulation and measure performances of the low-pass filter realized in SSL technology using RT/Duroid as substrate.

Fig 17 Variation of the resonator reflection coefficient as a function of frequency in the Ka band (L₀=4.45mm).

$|S_{11}|$

$|S_{21}|$

Fig 18 Variation of S11 (dB) and S21 (dB) as a function of the resonator length at F=28.5GHz
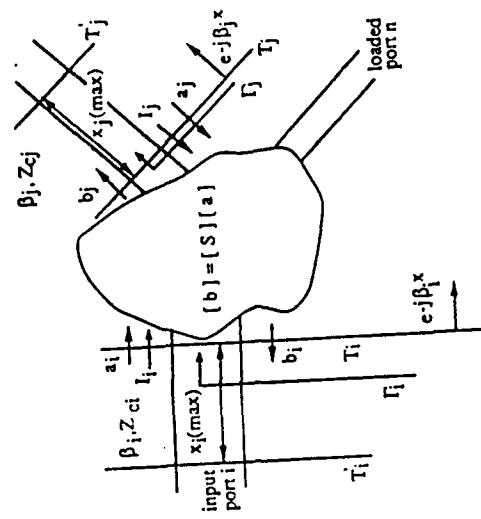
Fig. 25 General N-port discontinuity configuration

$$|S_{ii}| = \frac{-\alpha_{ii}}{1+\alpha_{ii}} + \frac{1}{2}\sqrt{\left(\frac{2\alpha_{ii}}{\alpha_{ii}+1}\right)^2 - 4\left(\frac{\alpha_{ii}-1}{\alpha_{ii}+1}\right)}$$

$$\alpha_{ji} = \sum_{\substack{j=1 \\ j\neq i}}^{N} k_{ji}$$

where

$$k_{ji} = \frac{|I_i(max)|^2}{|I_j(max)|^2}\frac{Z_{cj}}{Z_{ci}}\frac{1}{(1+|\Gamma_j|)^2}$$

$$\angle\Gamma_i - 2\beta_i x_i(max) = (2m+1)\pi, \quad \angle\Gamma_j + 2\beta_j x_j(max) = (2m+1)\pi, \quad m = 0,1,\dots,\infty$$

$$S_{ji} = \frac{-I_j(max)}{I_i(max)}\frac{\sqrt{Z_{cj}}}{\sqrt{Z_{ci}}}\frac{e^{j\beta_j x_j(max)}}{e^{-j\beta_i x_i(max)}}\frac{(1+|S_{ii}|)}{(1+|\Gamma_j|)}$$

$$\angle S_{ji} = \angle -I_j(max) - \angle I_i(max) + \beta_j x_j(max) + \beta_i x_i(max)$$

Fig.23 Current distribution along a loaded microstrip line
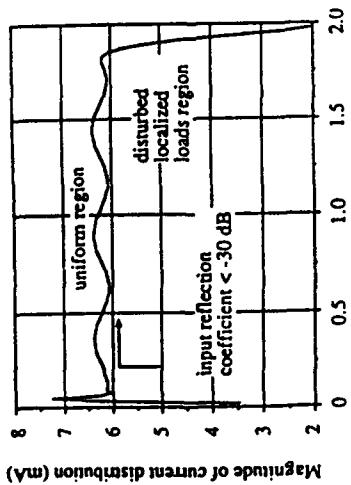
Fig. 24 Variation of the return loss as a function of frequency in the Ku-band for a matched load on a microstrip line (a=80 mm, b=7.112 mm, c=3.56 mm, w=0.75 mm, h=254 μm, εr=2.22)

Module S21 (dB)

Frequency(GHz)

Fig. 27 : Variation of |S21| (□Theory and △ measurements) for a
SSL stub discontinuity .

Modules

Frequency (GHz)

Fig.26 : Variation of S11 and S21 for a SSL bend discontinuityas
a function of frequency (W1=W2=W,a=b=12.7mm,c=3.81
mm, $\in$ r=9.8,h=0.635mm, ⊡ our analysis, △ reference (28))

[24] MATTHAEI G L, YOUNG L, JONES E M T, "Microwave filters, impedance matching networks and coupling structures", Mc Graw-Hill, New York, 1964.

[25] TEMES G C, MITRA S K, "Modem filter theory and design", John Wiley & Sons Ltd, USA, 1973.

[26] EDWARDS T C, "Foundations for microstrip circuit design", John Wiley & Sons Ltd, USA, 1981.

[27] EESOF Inc, Touchstone & Libra applications", USA, 1990.

[28] A. Hill and V.K. tripathi, "An efficient algorithm for the three-dimensionnal analysis of passive microstrip components and discontinuities for microwave and millimeter-wave integrated circuits " IEEE trans Microwave theory Tech, vol MTT-39, pp. 83-91, Jan. 1991.

[9] PICON O, "Traitements électromagnétiques 3-D applicables aux composants passifs micro-électroniques en ondes millimétriques", Thèse de Docteur ès-sciences, Rennes, 1988.

[10] ACHKAR J, PICON O, FOUAD HANNA V, CITERNE J, "Etude des modes d'une ligne asymétrique et de lignes couplées en technologie ligne microruban suspendu par la méthode d'approche spectrale", 10ème Colloque Optique Hertzienne et Diélectrique, Rennes, pp 247-250, 1989.

[11] ACHKAR J, PICON O, FOUAD HANNA V, CITERNE J, "Analysis of symmetric and asymmetric coupled suspended striplines and some associated discontinuities", 20th European Microwave Conference, Budapest, pp 537-542, 1990.

[12] OMAR A S, SCHUNEMANN K F, "Formulation of the singular integral equation technique for planar transmission lines", IEEE Trans on Microwave Theory and Techniques, Vol MTT-33, pp 1313-1322, 1985.

[13] GARDIOL F, "Hyperfréquences", Traité d'électricité, Vol 13, Dunod, Paris, 1987.

[14] YUNYI W, KAIJUN G, YONGHUI S, "Synthesis equations for shielded suspended substrate microstrip line and broadside coupled stripline", IEEE MTT-S, USA, pp 331-334, 1988.

[15] KOUL S K, BHAT B, "Propagation parameters of coupled microstrip like transmission lines for millimeter wave applications", IEEE Trans on Microwave Theory and Techniques, Vol MTT-29, pp 1364-1369, 1971.

[16] COLLIN R E, "Field theory of guided waves", Mc Graw-Hill, New York, 1960.

[17] ACHKAR J, "Etude et caractérisation de lignes microruban suspendu en vue de la réalisation de composants hyperfréquences passifs", Thèse de Doctorat, Paris, 1991.

[18] HARLAN HOWE Jr, "Stripline circuit design", Artech house, Massachusetts, 1974.

[19] HAMMERSTAD E, JANSEN R H, "Accurate models for microstrip computer aided design", IEEE International Microwave Symposium, pp 407-409, 1980.

[20] PICON O, ACHKAR J, "Logiciels de calcul de caractéristiques de lignes planaires fermées et de discontinuités uniaxiales associées", Journées d'Etudes sur la CAO Hyperfréquences, Chapitre IEEE MTT-S, Poigny-La-Forêt, 1989.

[21] COHN S B, "Shielded coupled-strip transmission line", IEEE Trans on Microwave Theory and Techniques, Vol MTT-3, pp 29-38, 1955.

[22] ACHKAR J, PICON O, FOUAD HANNA V, CITERNE J, "Coupleur directif en technologie ligne microruban suspendu dans le domaine millimétrique", Annales des Télécommunications, Vol 45, N° 5-6, pp 296-299, 1990.

[23] ACHKAR J, PICON O, FOUAD HANNA V, CITERNE J, Theoretical and experimental investigation of some different resonators in suspended stripline technology", 21th European Microwave Conference, Stuttgart, pp 1041-1046, 1991.

13

12

Module

REFERENCES

[1]   SPIELMAN B E, "Dissipation loss effects in isolated and coupled transmission lines", IEEE Trans on Microwave Theory and Techniques, Vol MTT-16, pp 648-656, 1977.

[2]   PICON O, ACHKAR J, FOUAD HANNA V, CITERNE J, "Scattering parameters of a step discontinuity in a suspended microstrip line", Microwave and Optical Technology Letters, USA, Vol 1, N°7, pp 240-243, 1988.

[3]   ACHKAR J, PICON O, FOUAD HANNA V, CITERNE J, "Frequency dependent characteristics of an asymmetric suspended stripline and its open end discontinuity for millimeter wave applications", 19th European Microwave Conference, London, pp 767-772, 1989.

[4]   ITOH T, "An overview on numerical techniques for modelling 3-dimensional passive components", 15th European Microwave Conference, Paris, pp 1059-1063, 1985.

[5]   ITOH T, "Overview of quasi-planar transmission lines", IEEE Trans on Microwave Theory and Techniques, Vol MTT-37, pp 275-280, 1989.

[6]   WEXLER A, "Solution of waveguide discontinuities by modal analysis", IEEE Trans on Microwave Theory and Techniques, Vol MTT-15, pp 508-517, 1967.

[7]   MITTRA R, ITOH T, "A new technique for the analysis of dispersion characteristics of microstrip lines, IEEE Trans on Microwave Theory and Techniques, Vol MTT-19, pp 47-56, 1971.

[8]   GUPTA K C, GARG R, BAHL I J, "Microstrip lines and slotlines", Artech house, Massachusetts, 1979.

Fig. 28: Variation of S21,S11 (dB) for a bent-stub discontinuity as a function of frequency (L1=2.358mm, L2=1.572mm, W=0.786mm).

LEAKAGE AND CROSSTALK EFFECTS IN MILLIMETER - WAVE INTEGRATED CIRCUITS

A. A. Oliner

Polytechnic University • USA

LEAKAGE AND CROSSTALK EFFECTS
IN MILLIMETER-WAVE INTEGRATED CIRCUITS

Arthur A. Oliner

Professor Emeritus
Polytechnic University
Brooklyn, New York, USA

New Address: 11 Dawes Road
                Lexington, MA 02173
                USA

Telephone: 617-863-5909
Fax: (same)

INTRODUCTION

It is evident that crosstalk in microwave and millimeter-wave integrated circuits can seriously disturb and even ruin circuit performance. It is therefore important to understand how crosstalk can be produced and what circuit effects can result from it.

Unwanted coupling between two different portions of a circuit can occur when two lines, for example, are too closely spaced to each other, so that one is in the near field of the other. This type of crosstalk has been well understood for a long time, and it will not be discussed here. We will examine, instead, the crosstalk that results from the excitation of *surface waves* within the circuit. These surface waves can travel throughout the circuit, bounce around, and cause various effects, ranging from direct interaction to package resonances.

Surface waves can be excited by two principal mechanisms:

    (a) circuit discontinuities, and

    (b) power leakage from the transmission-line dominant mode.

The fact that surface waves are excited at transmission-line discontinuities is of course well known, but what effects these surface waves produce have been examined only infrequently. We will discuss here the conditions under which discontinuities can interact with other portions of an integrated circuit, and how discontinuities can excite package guided modes and package cavity resonances.

The fact that *transmission-line dominant modes* can leak became known only a few years ago, principally as a result of work by Shigesawa, Tsuji and Oliner. The leakage occurs in the form of a surface wave that propagates away at an angle from the axis of the transmission line. For most printed-circuit transmission lines, the dominant mode becomes leaky only at the higher frequencies, so that leakage effects become significant principally at millimeter wavelengths. The talk will discuss precisely when a dominant mode can become leaky, how the transmission line, when it leaks, can interact with other lines and with discontinuities, and how such leakage can excite package guided modes (whereas the dominant mode cannot interact with package guided modes when it is purely bound).

The present talk will begin with some necessary background material involving various constituent modes and their dispersion properties. The three printed-circuit lines that we consider are microstrip line, slot line and coplanar waveguide. We next discuss the way power can leak from the dominant mode on printed-circuit lines. In that connection, we first establish when a mode can

become leaky, then describe the physical properties of the leakage, and finally present several examples of such leakage. After that, we examine the effects of crosstalk due to such leakage and also to transmission-line discontinuities, along the lines indicated earlier. Finally we look at package effects, and examine how discontinuities and leakage may each influence package guided modes and participate in resonances of the package as a whole.

The stress in the presentation will be on the physical principles and the performance consequences.

OUTLINE OF TALK

I. BACKGROUND AND REVIEW

  A. Constituent Basic Modes on Printed-Circuit Lines

  B. Dispersion Properties

    1. Modes on printed-circuit lines
    2. Surface waves

II. POWER LEAKAGE FROM PRINTED-CIRCUIT LINES

  A. When Can a Mode Become Leaky?

  B. Physical Properties of the Leakage

  C. Examples of Leakage

III. SOURCES OF CROSSTALK

  A. Waveguide Discontinuities

  B. Leakage Effects

IV. PACKAGE EFFECTS

  A. Guided Modes and Cavity Modes

  B. Coupling Between Leaky Modes and Guided Modes

  C. Excitation by Discontinuities

CONSTITUENT BASIC MODES

1. MODES ON PRINTED - CIRCUIT LINES

Microstrip Line
Slot Line
Coplanar Waveguide (CPW)

2. SURFACE WAVES ON THE SURROUNDING SUBSTRATE

(They are different when the line does or does not have a ground plane.)

ϵ_r ϵ_r

3. MODES OF THE PACKAGE

Guided Modes and Cavity Modes
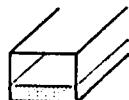
waveguide   resonant cavity

5

DISPERSION PROPERTIES
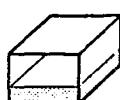
<u>wavenumber</u> $= \beta - j\alpha$

β = phase constant
α = attenuation constant

α is due to material losses, and to leakage losses, where appropriate.

If only leakage is included, then α is called the leakage constant.

<u>Dispersion</u> means the variation of wavenumber with frequency.

Customary to <u>plot</u> as β/k_0 vs. frequency (or in normalized form, such as h/λ_0).

$k_0 = 2\pi/\lambda_0$ = free-space wavenumber

$$\frac{\beta}{k_0} = \frac{\lambda_0}{\lambda_g} = \frac{c}{v_p}$$

c = velocity of light in free space
v_p = phase velocity of the mode

$$\epsilon_{eff} = \left(\frac{\beta}{k_0}\right)^2 ; \quad n_{eff} = \frac{\beta}{k_0}$$

DISPERSION BEHAVIOR OF MODES ON PRINTED - CIRCUIT LINES

1. Microstrip Line

OC

somewhat dispersive

2. Slot Line

SC
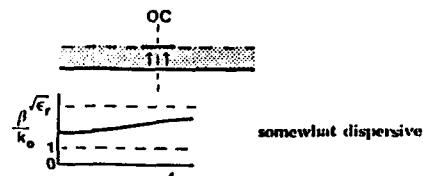
less dispersive
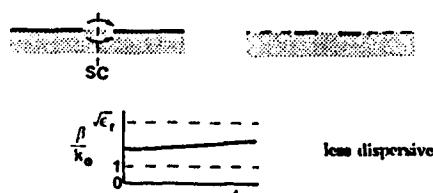
3. Coplanar Waveguide (CPW)

OC

least dispersive
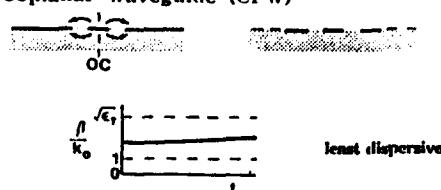
DISPERSION BEHAVIOR OF SURFACE WAVES

1. Ungrounded Layer

$\sqrt{\epsilon_r}$
$\frac{k_z}{k_0}$
TE_0 TM_0 TE_1

f (or h/λ_0)

$k_z = \beta$ of the surface wave
Strongly dispersive at higher frequencies

2. Grounded Layer

$\sqrt{\epsilon_r}$
$\frac{k_z}{k_0}$
TM_0 TE_1

f
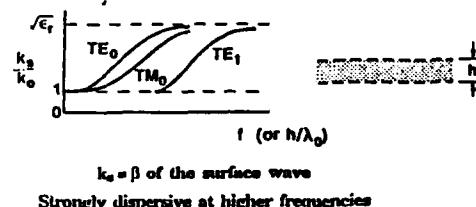
EXAMPLES

grounded   grounded   ungrounded
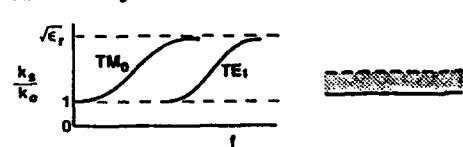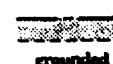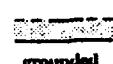
If conductor - backed

$\frac{k_{PP}}{k_0}$ $\sqrt{\epsilon_r}$

parallel - plate mode   grounded

7
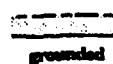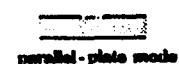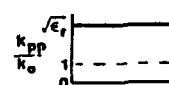
WHEN CAN A DOMINANT MODE BECOME LEAKY?

Suppose there is leakage:

$$\cos\theta = \beta/k_o$$
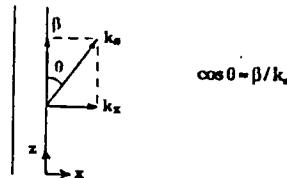
Therefore, for real power to leak away we must have

$$\beta < k_o$$

or

$$\beta/k_o < k_o/k_o$$

Example: CPW of finite width

bound ← → leaks
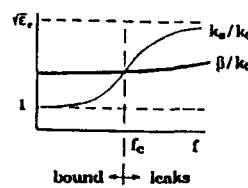
For $f > f_c$, the dominant mode on the CPW leaks.

DISPERSION PLOTS FOR VARIOUS CASES

1. <u>Conductor-backed slot line or CPW of infinite width</u>

$\frac{\beta}{k_o} < \frac{k_{PP}}{k_o}$; leakage at <u>all</u> frequencies in parallel-plate guide form
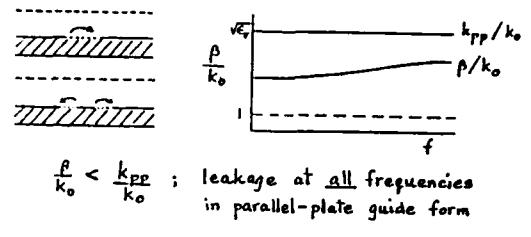
2. <u>Conductor-backed slot line or CPW of finite width</u>
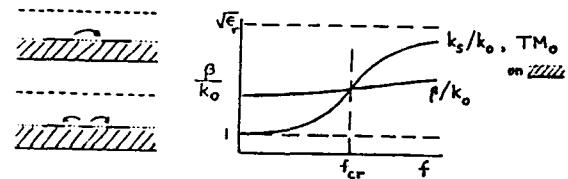
At low frequencies, $\frac{\beta}{k_o} > \frac{k_s}{k_o}$; bound mode

Above f_{cr} , $\frac{\beta}{k_o} < \frac{k_s}{k_o}$; leakage in surface wave form, TM_0 on ▨▨

9

3. <u>Conventional slot line or CPW of infinite width</u>
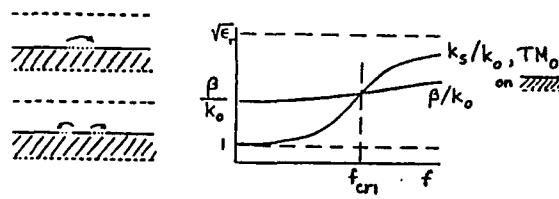
At low frequencies, $\frac{\beta}{k_o} > \frac{k_s}{k_o}$; bound wave

Above f_{cr1} , $\frac{\beta}{k_o} < \frac{k_s}{k_o}$; leakage in surface wave form, TM_0 on ▨▨

4. <u>Conventional slot line or CPW of finite width</u>
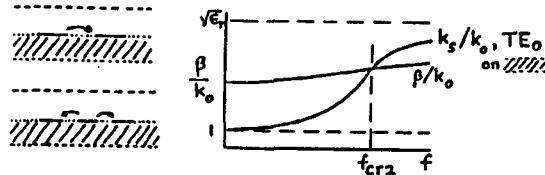
At low frequencies, $\frac{\beta}{k_o} > \frac{k_s}{k_o}$; bound wave

Above f_{cr2} , $\frac{\beta}{k_o} < \frac{k_s}{k_o}$; leakage in surface wave form, TE_0 on ▨▨

NORMALIZED PHASE AND LEAKAGE CONSTANTS OF CONDUCTOR-BACKED SLOT LINE

11
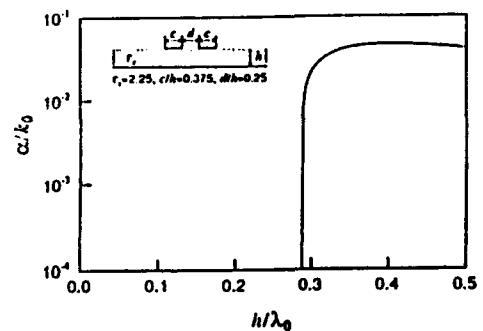
NORMALIZED PHASE AND LEAKAGE CONSTANTS FOR MICROSTRIP LINE ON AN EPSILAM 10 SUBSTRATE
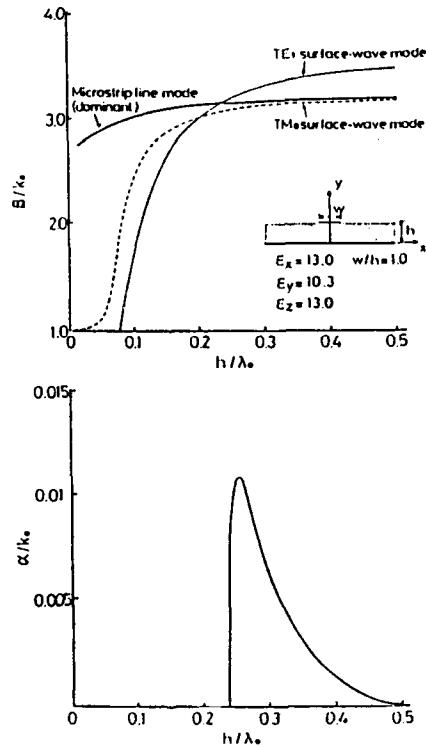
LEAKAGE FROM HIGHER MODES ON MICROSTRIP LINE
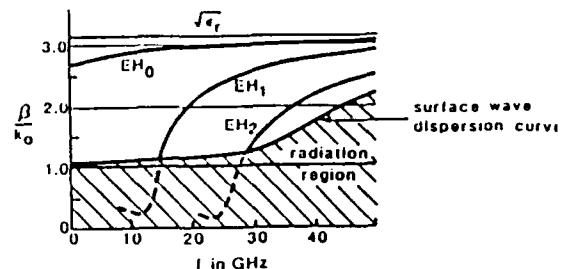
The solutions are leaky where the curves are shown dashed. In that region just above cutoff, the surface wave leaks over a wide range of angles, from endfire to almost broadside as the frequency decreases toward cutoff.

If there is no top cover, leakage also occurs into space waves when $\beta / k_0 < 1$.

SOURCES OF CROSSTALK

1. Waveguide Discontinuities

2. Leakage Effects

WAVEGUIDE DISCONTINUITIES

What occurs physically?

Top view of microstrip
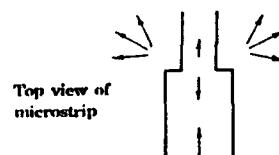
1. Higher modes on the line are excited, but decay away from it, and their power is essentially stored in the neighborhood of the discontinuity. They provide reactance to the equivalent network of the discontinuity.

2. The lowest surface wave (or waves) on the surrounding substrate is excited. It radiates power in surface - wave form over a wide range of angles, and it provides resistance to the equivalent network of the discontinuity.

3. If there are additional dominant modes, there will be some mode conversion into them. This effect contributes to resistance.

COUPLING BETWEEN "LEAKY" MODES AND PACKAGE GUIDED MODES

1. We can view the package guided mode in terms of a pair of surface waves at an angle.

Top view $\beta_p = k_s \cos \theta_p$
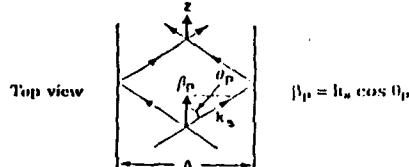
2. The package guided mode can be excited by the "leaky" mode when the leakage angle θ becomes the same as the package guided mode angle θ_p. The two modes then couple to each other.

Top view

We had:
$$\beta = k_s \cos \theta$$

Above:
$$\beta_p = k_s \cos \theta_p$$

Thus: $\beta = \beta_p$

when $\theta = \theta_p$
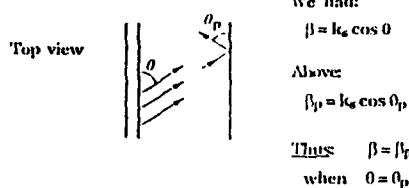
COUPLING BETWEEN "LEAKY" MODES AND PACKAGE GUIDED MODES
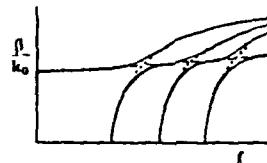
The dashed and dotted lines represent the original (unperturbed) transmission line mode and package guided modes, respectively. The solid lines represent the coupled mode result.

FURTHER INFORMATION

Package guided mode
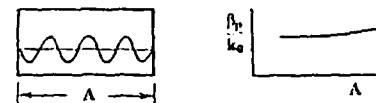
"Leaky" mode
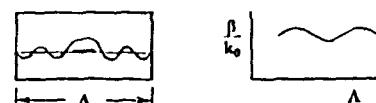
Bound mode
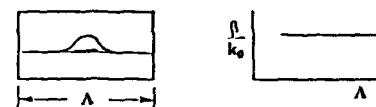
EXCITATION OF PACKAGE GUIDED MODES BY DISCONTINUITIES
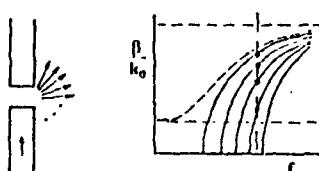
1. The discontinuity excites surface waves at all angles, so that their β values range from 0 to k_s.

2. Therefore, at a particular frequency, they can couple to all package guided modes with β values in that range, usually a large number of them.

3. In the above example, 4 package guided modes are excited.

4. By contrast, a transmission line mode, with a given β at frequency f, could at most excite (couple to) a single package guided mode, and perhaps none unless the frequency were shifted.
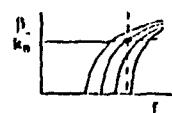
EXCITATION OF PACKAGE CAVITY RESONANCES BY DISCONTINUITIES
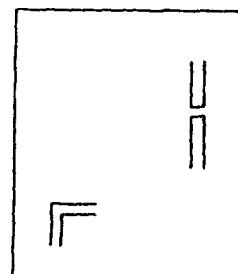
These two discontinuities can interact via the surface wave supported by the substrate. The coupling is ordinarily quite small, but gets large at frequencies corresponding to package cavity resonances. For a large package, the resonances become closely spaced in frequency.

The cavity Q, and therefore the undesired coupling or crosstalk, can be reduced by introducing loss, but the resonances broaden and can form almost a continuum in frequency in large packages.

CROSSTALK BETWEEN LINES AND DISCONTINUITIES

Is there crosstalk?

NO
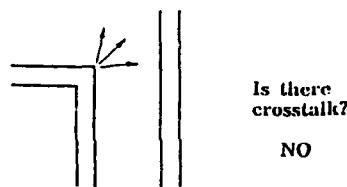
(The surface waves have wavenumbers that vary from k_o to 0 (or negative) in the component along the line length. They will couple only if the k_o component is $= \beta$. Here $\beta > k_o$.)

Is there crosstalk?

YES

CROSSTALK DUE TO LEAKAGE

1. Two Identical Lines

Do they couple?

YES

(Directional coupler action, at a distance.)

2. Two Different Lines

Do they couple?

NO

(α and β must be the the same for coupling)
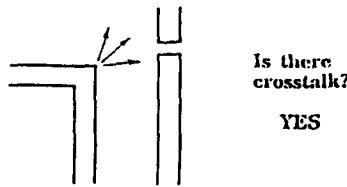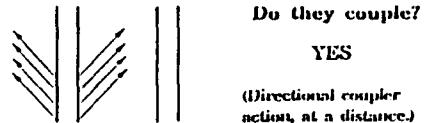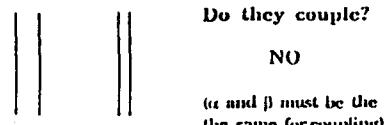
3. If There Is a Discontinuity on the Second Line

Crosstalk?

YES

17

PACKAGE MODES

Two types: Guided modes
Cavity modes

EXAMPLE OF GUIDED MODES
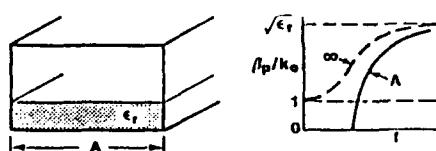
If metal bounding walls:

$$\beta_p^2 = \beta_c^2 - (n\pi/A)^2$$
$$\beta_\infty = k_o$$

Thus, there is a simple relation between the package guided modes and surface waves on an unbounded substrate. There are many package guided modes for each possible surface wave. If A is large, they are closely spaced in wavenumber.

WHAT HAPPENS TO LEAKAGE WHEN THE PACKAGE IS FINITE?

Infinite width

Finite width
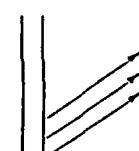
Complex wavenumber
$\beta - j\alpha$

Real wavenumber
β

The "leakage" power is reflected from the boundaries, but it is still spread throughout the circuit, and still can cause crosstalk.

3D MODELING OF MILLIMETER - WAVE CIRCUITS

AND ANTENNAS

N. Alexopoulos

UCLA • USA

Unified Theory for

Printed Circuits, Antennas

and
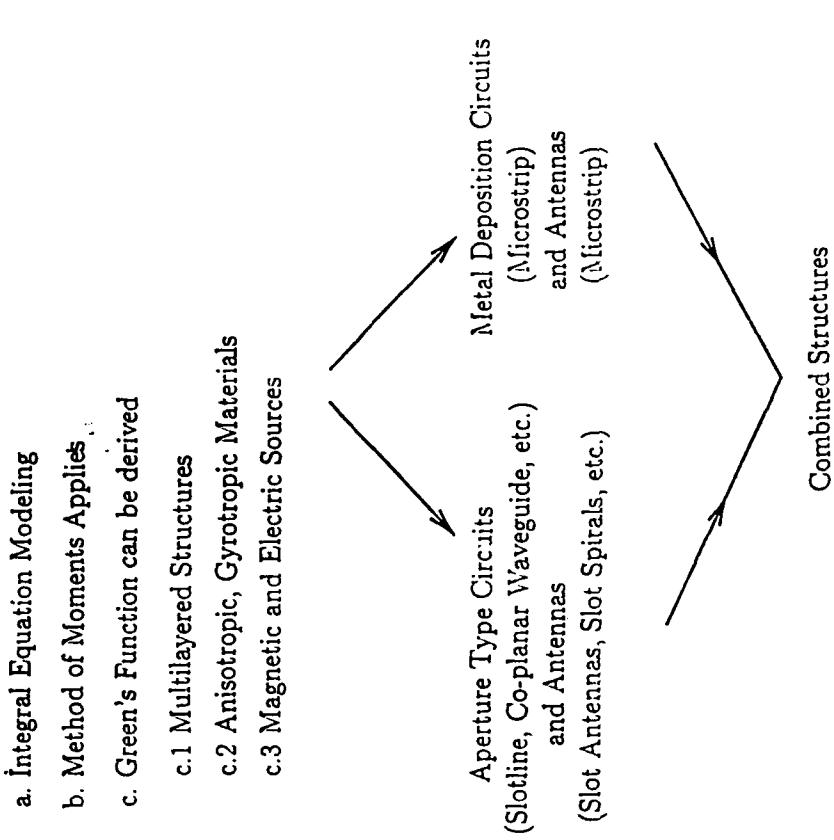
Scatterers

N.G. Alexopoulos

Electrical Engineering Department

University of California

Los Angeles

CA 90024

List of Some Advantages of this Philosophy

a. Integral Equation Modeling

b. Method of Moments Applies

c. Green's Function can be derived

   c.1 Multilayered Structures

   c.2 Anisotropic, Gyrotropic Materials

   c.3 Magnetic and Electric Sources

Metal Deposition Circuits
(Microstrip)
and Antennas
(Microstrip)

Aperture Type Circuits
(Slotline, Co-planar Waveguide, etc.)
and Antennas
(Slot Antennas, Slot Spirals, etc.)

Combined Structures

Fig. Balanced mixer

via hole

coplanar waveguide

slotline

LO

IF

microstrip

diodes

microstrip filter

RF

Microstrip Circuits - Aperture Antennas and Scatterers

ε_n ••• ε_3 ε_2 ε_1

ε_n, μ_n ••• ε_3, μ_3 ε_2, μ_2 ε_1, μ_1

Microstrip Circuits, Antennas and Scatterers

ε_n ••• ε_3 ε_2 ε_1

ε_n, μ_n ••• ε_3, μ_3 ε_2, μ_2 ε_1, μ_1
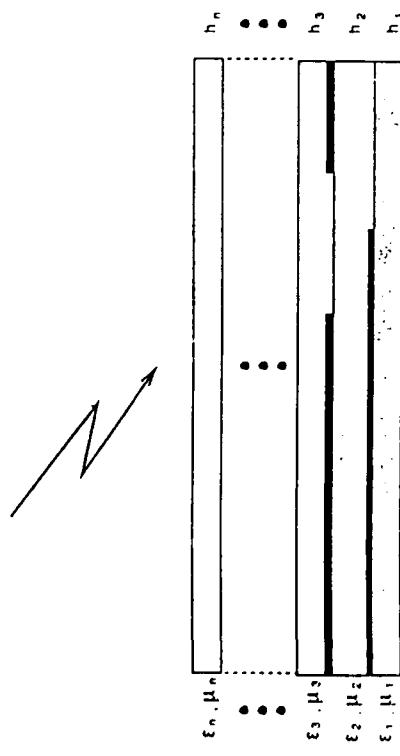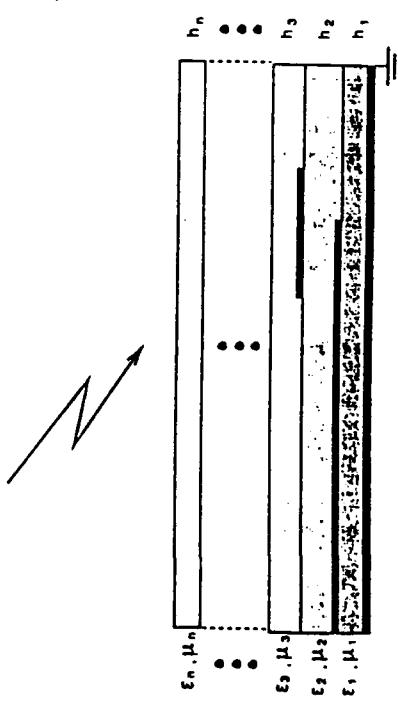
Fig. Linear array of microstrip dipoles

Dipoles

Fig Microstrip corporate feed

Incident wave

Feed

Microstrip Linear Array of EMC Dipoles
with a Corporate Feed

Tzyy-Sheng Horng * and Nicolaos G. Alexopoulos

Electrical Engineering Department

University of California, Los Angeles

Los Angeles, CA90024

June, 1991

6
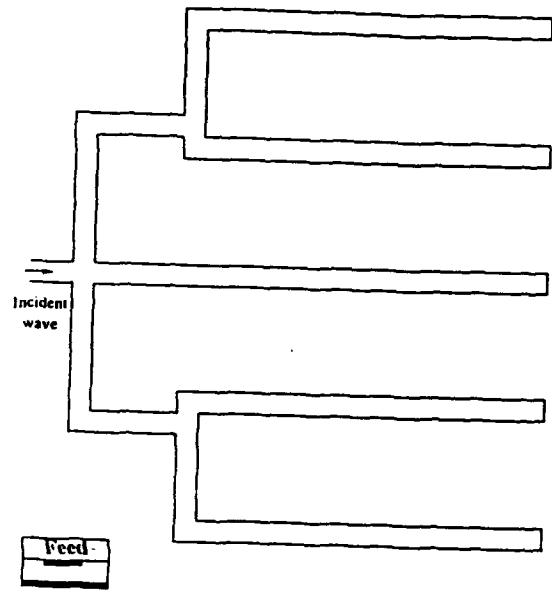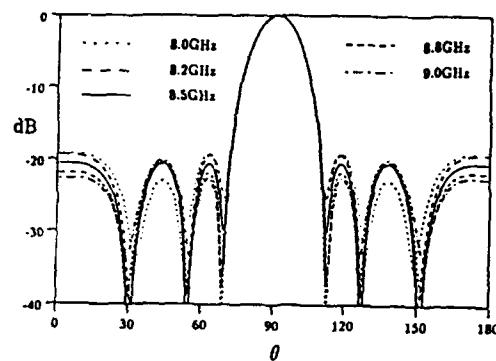
Fig. 7. Array pattern
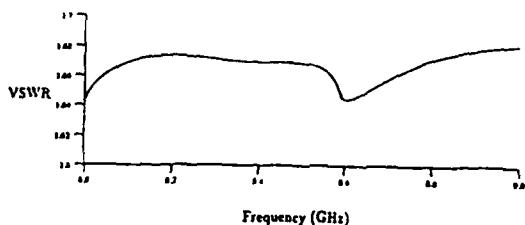
Fig. 8. VSWR versus Frequency

$\varepsilon = 2.35$

fo ≈ 8.5 GHz
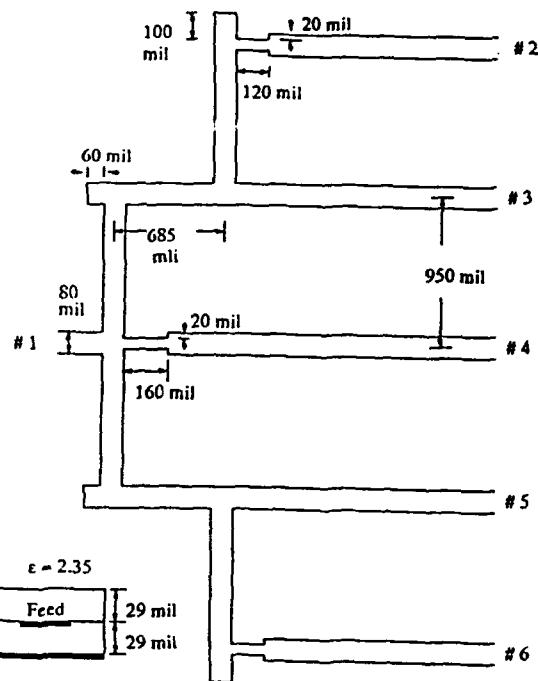
Fig. 6. An embedded microstrip corporate feed

Ld = 442 mil, Od = 15 mil

$\varepsilon = 2.35$
$\varepsilon = 2.35$
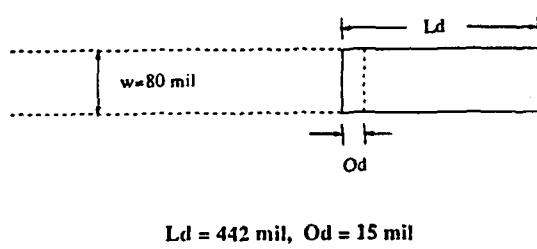
h2 = 29 mil
h1 = 29 mil
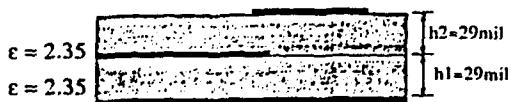
Fig. Electromagnetically coupled dipole fed by a microstripline
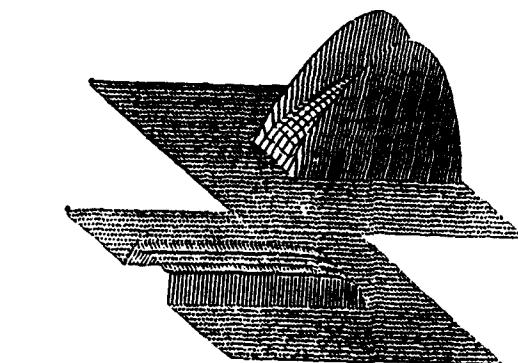
Fig. The magnitude of microstrip currents in an isolated EMC dipole
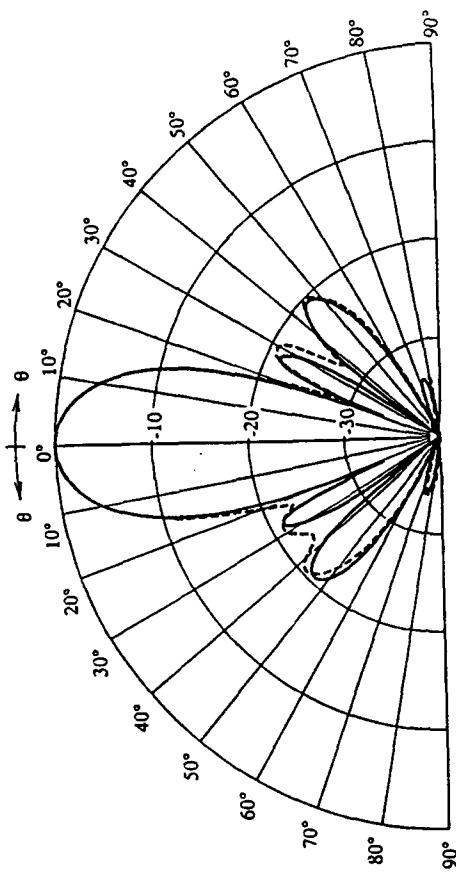
H Plane Pattern for a Linear 5 Element Array Microstrip Dipoles
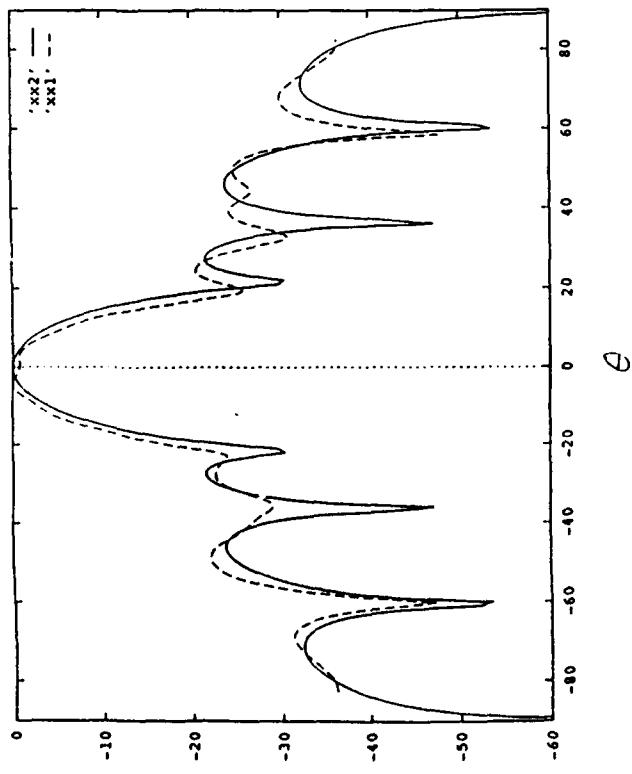
Phase Image of
5 Element Dipole Array

file = jason1.dat
freq = 8.5 GHz

x (wavelengths)

y (wavelengths)

Levels

0 -> 30
30 -> 60
60 -> 90
90 -> 120
120 -> 150
> 150

Levels

150
120 > 150
90 < 120
60 < 90
-30 < 0

Microstrip Linear Array of EMC Dipoles with a Corporate Feed

- 440 mil
- 20 mil
- 945 mil
- 0.275, -21.00
- 612 mil
- 0.444, -22.38
- 4, -23.34
- 0.532, -24.318
- = 1.0
- = 2.35   29 mil
- = 2.35   29 mil
- 0.444, -22.38
- f o = 8.5 GHz
- 0.275, -22.34
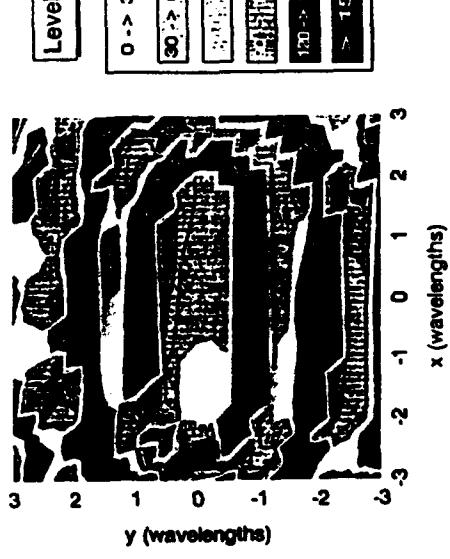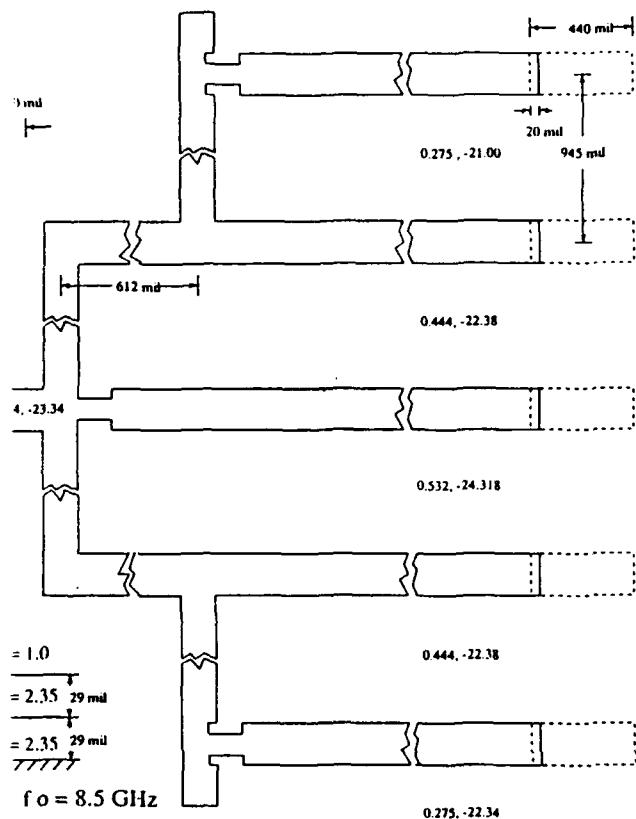
- Relative current distribution in the output ports

  1∠3.40°   2.42∠1.22°   3.15∠0°   2.42∠1.22°   1∠3.40°

  - VSWR   1.67
  - Directivity   5.77
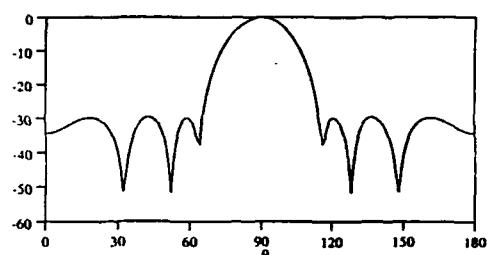
Fig. Array pattern

Phase Image of
5 Element Dipole Array

file = jason1.dat
freq = 8.5 GHz

Levels

0 -> 30   30 -> 60   ...   > 150

x (wavelengths)

y (wavelengths)
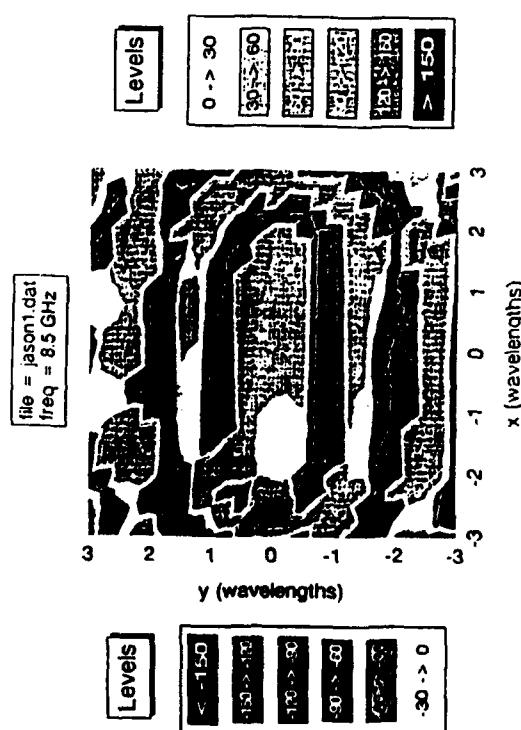
Magnitude Image of
5 Element Dipole Array

file = jason1.dat
freq = 8.5 GHz

Levels

< 2.0   2.0 -> 2.5   2.5 -> 3.0   ...   > 4.5

x (wavelengths)

y (wavelengths)
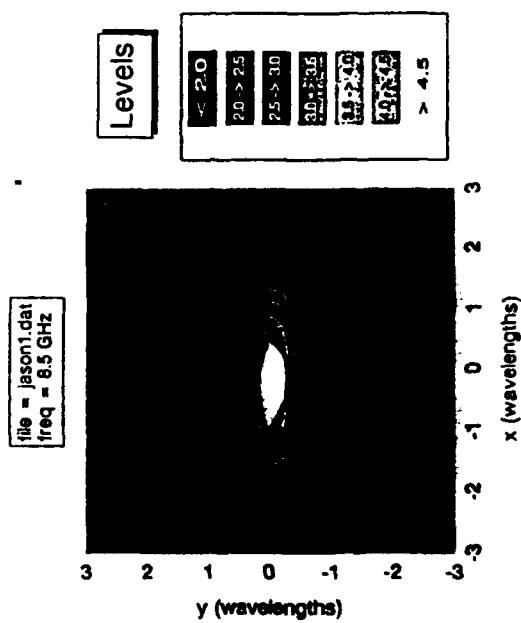
• Computation Time

GOAL: To calculate the matrix elements $Z_{ij}(m,n)$ between any two basis functions

where

i,j identify type of basis function

$$i,j = \begin{cases} 1, & \text{Rectangular subdomains} \\ 2, & \text{Triangular subdomains} \\ 3, & \text{Semi-infinite traveling wave modes} \end{cases}$$

m,n are relative distance indices, $m = [-50:50], n = [-50:50]$

Computation time on IBM-3090 for matrix elements

- Z_{11}: 2100 sec   *2 rect subdomains*   *setting up data base*

- Z_{22}: 3900 sec   *2 triang subdomain*

- Z_{13}: 8500 sec

Triangular basis function
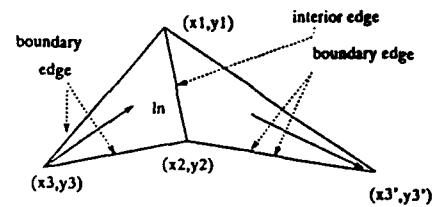
Ref : Rao, Wilton and Glisson, AP-30, pp.409-418, May 1982

For each interior edge l_n, the triangular basis function is defined as

$$\vec{F}_n(x,y) = \begin{cases} \frac{l_n}{2A_n^+}(x-x3,\,y-y3), & x,y \text{ in } T_n^+ \\ \frac{l_n}{2A_n^-}(x3'-x,\,y3'-y), & x,y \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases}$$

• Method of Moments

Basis Functions

$$\vec{J}(x,y) = \sum_n C_n^e \vec{J}_n(x,y)$$

$$\vec{M}(x,y) = \sum_n C_n^m \vec{M}_n(x,y)$$

Integral Equations

$$< \vec{E}(x,y) \cdot \vec{J}_n(x,y) > = 0, \quad n = 1,2,\cdots,N, \quad x,y \text{ on microstrips}$$

$$< \Delta\vec{H}(x,y) \cdot \vec{M}_n(x,y) > = 0, \quad n = 1,2,\cdots,N, \quad x,y \text{ on slots}$$

Matrix Form

$$[Z]_{N \times N} [I]_N = [V]_N$$

~ 5000 unknowns
~ 600 Mbytes memory
~ 40 minutes
IBM 3090

Matrix Element

$$Z_{ij}(m,n) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \tilde{G}_{ij}(k_x,k_y) \tilde{J}_i^{m^*}(k_x,k_y) \tilde{J}_j^n(k_x,k_y) dk_x dk_y$$

IMSL package

• Expansion Functions

1. Rectangular subdomain modes

$$\vec{J}_n, \vec{M}_n = \begin{cases} PWS(x)P(y)\hat{x}, & x\text{-directed currents} \\ PWS(y)P(x)\hat{y}, & y\text{-directed currents} \end{cases}$$

2. Triangular subdomain modes

$$\vec{J}_n, \vec{M}_n = \begin{cases} \frac{l_n}{2A_n^+}(x-x_3,\,y-y_3), & x,y \text{ in } T_n^+ \\ \frac{l_n}{2A_n^-}(x_3'-x,\,y_3'-y), & x,y \text{ in } T_n^- \\ 0, & \text{otherwise} \end{cases}$$

3. Semi-infinite traveling wave modes (entire domain modes)

$$\vec{J}_n, \vec{M}_n = \begin{cases} U(x-x_r)e^{\pm j\beta(x-x_r)}P(y)\hat{x}, & x\text{-directed currents} \\ U(y-y_r)e^{\pm j\beta(y-y_r)}P(x)\hat{y}, & y\text{-directed currents} \end{cases}$$

• Testing Functions

Subdomain modes only

Fig. 1 A generic structure of a four port microstrip discontinuity

- Spectral-Domain Dyadic Green's Function
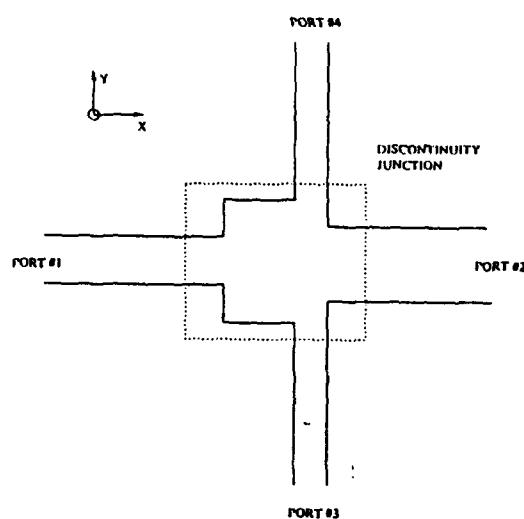
$$\begin{bmatrix} \vec{E}(x,y) \\ \triangle \vec{H}(x,y) \end{bmatrix} = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \begin{bmatrix} \overleftrightarrow{G}_{ee}(k_x,k_y) & \overleftrightarrow{G}_{em}(k_x,k_y) \\ \overleftrightarrow{G}_{me}(k_x,k_y) & \overleftrightarrow{G}_{mm}(k_x,k_y) \end{bmatrix}$$
$$\cdot \begin{bmatrix} \vec{\tilde{J}}(k_x,k_y) \\ \vec{\tilde{M}}(k_x,k_y) \end{bmatrix} e^{-jk_x x} e^{-jk_y y} dk_x dk_y$$

where

$$\tilde{F}(k_x,k_y) = \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} F(x,y) e^{jk_x x} e^{jk_y y} dx, dy$$
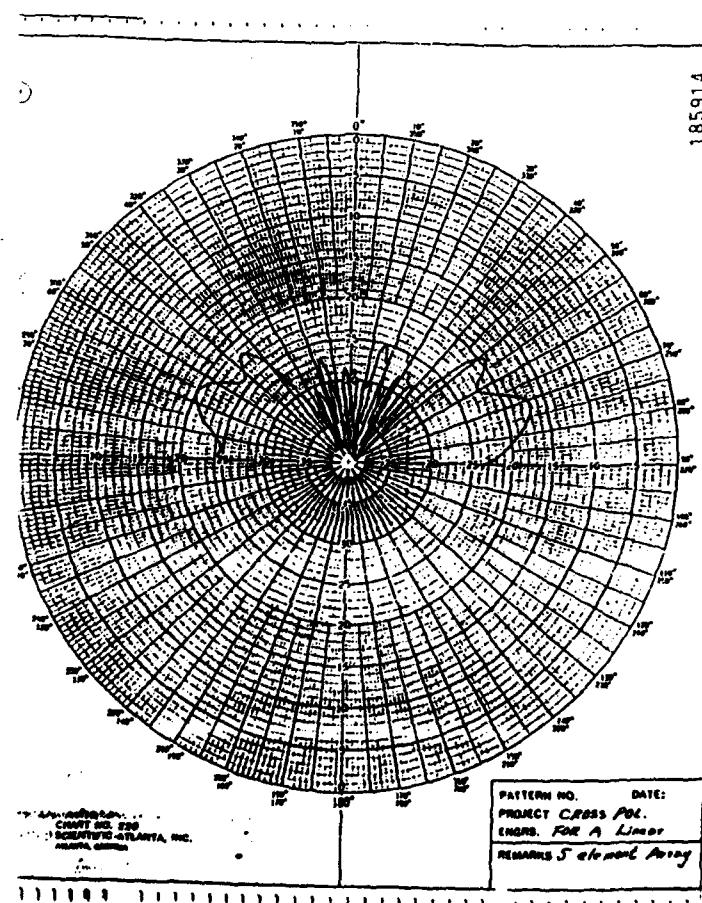
DISCONTINUITIES

JUNCTIONS

AND

TRANSITIONS
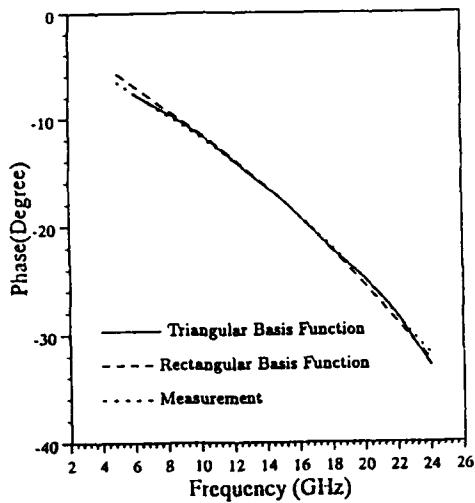
Fig. Phase of S parpmeters of an open-end
(ϵ_r=9.9, w=24mil, h=25mil)
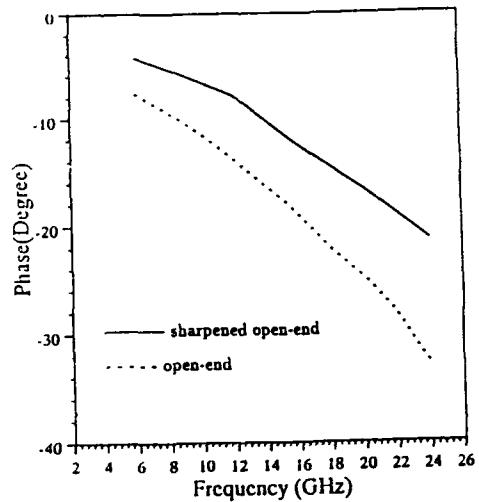
Fig. Phase of S parpmeters of a sharpened open-end and
an open-end (ϵ_r=9.9, w=24mil, h=25mil)

Basic Features Using Triangular Subdomains

1. Transmission lines have finite length.

2. Ideal current source is used for excitation.

3. The grid of subdomains is translated in the x,y plane to facilitate the
   computation of reaction elements.

4. Subdomains which are too distant from each other to be included in
   the grid have reactions which are set to zero.
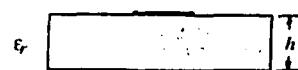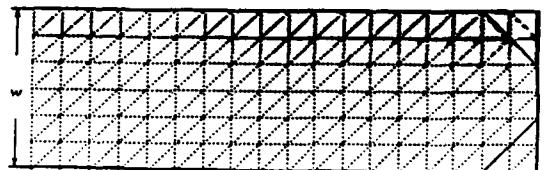
Fig. The geometry of an open-end and a sharpened open-end
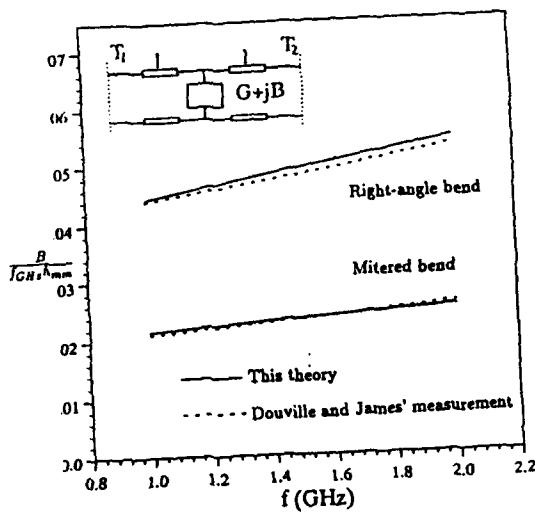
Fig. Frequency dependence of normalized susceptance for right-angle and mitered bend *discontinuities* (ϵ_r=10.8, w=4.572mm, h=5.08mm)
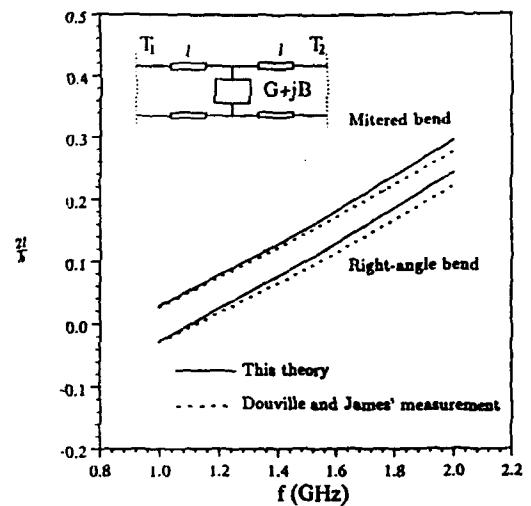
Fig. Frequency dependence of normalized electric length *for right-angle bend and mitered bend discontinuities* (ϵ_r=10.8, w=4.572mm, h=5.08mm)

$\epsilon_r = 9.9, w = 24mil, h = 25mil, f = 24GHz$

x directed current

y directed current
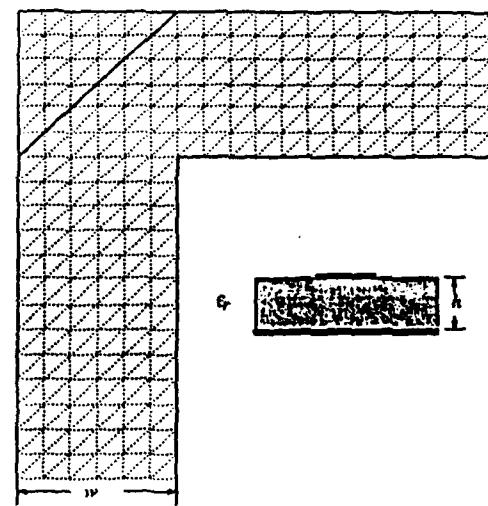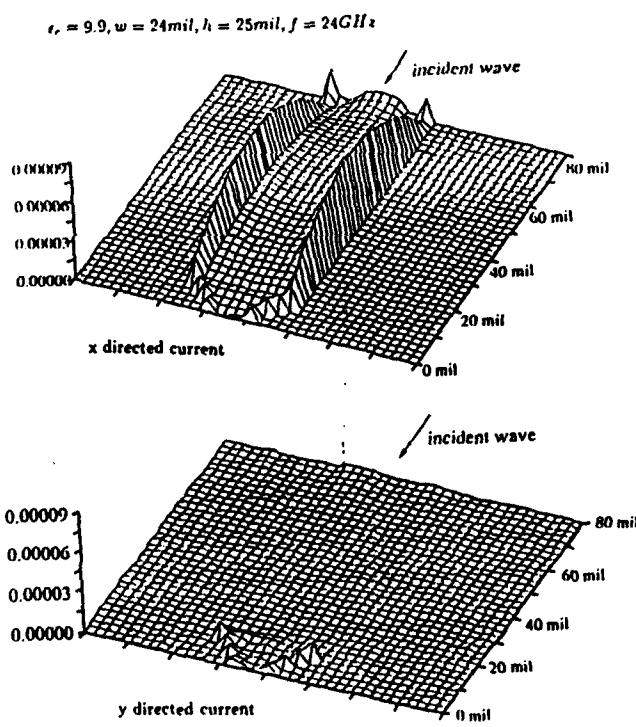
Fig. The geometry of right-angle bend and mitered bend discontinuities
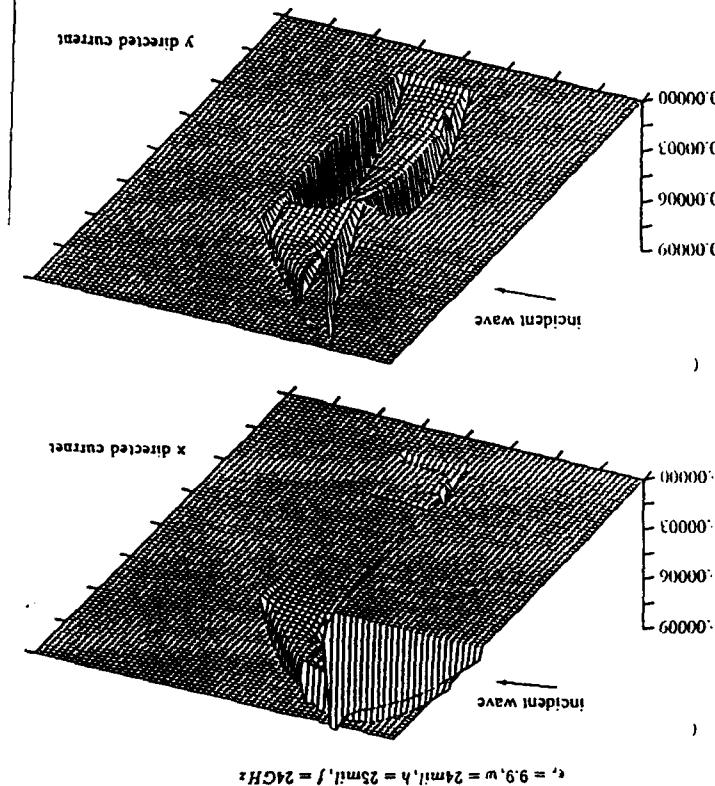
Fig. Phase of S33 of a basic tee junction
($\epsilon_r \approx 9.9$, w=24mil, h=25mil)
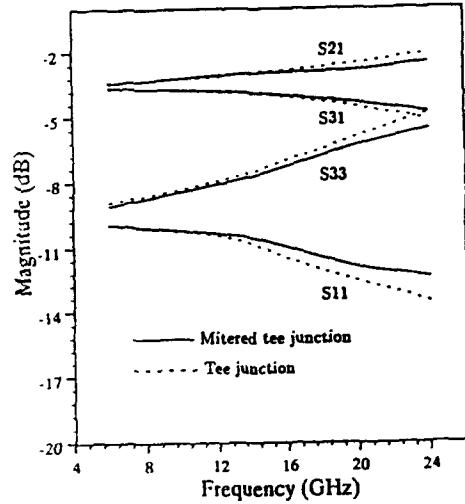
Fig. Magnitude of S parameters of tee and mitered tee junctions
($\epsilon_r = 9.9$, w=24mil, h=25mil)
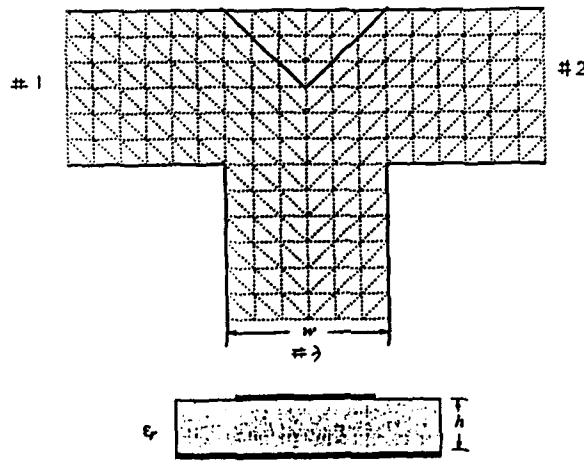
Fig. The geometry of tee and mitered tee junctions
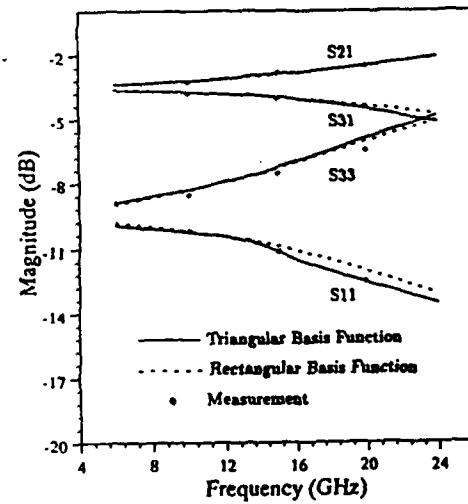
Fig. Magnitude of S parameters of a basic tee junction
($\epsilon_r = 9.9$, w=24mil, h=25mil)

Frequency (GHz)	Total Power	Reflected Power	Transmitted Power	Radiation Loss	Surface-wave Loss
20	1.001	0.134	0.822	0.022	0.023
21	0.997	0.153	0.787	0.027	0.030
22	1.001	0.172	0.755	0.033	0.041
23	1.005	0.193	0.719	0.039	0.054
24	1.007	0.208	0.678	0.048	0.073

Table 1. Power conservation check for right-angle bend discontinuity.
(Parameters are the same as those in Fig. 4.)

$$C_1 \cdot Z_o = \frac{\lambda}{2S}\left[-0.3154 + 1.1251\tanh\left(1.3576\frac{W}{h}\right)\right](pF \cdot \Omega) \quad (1)$$

$$C_2 \cdot Z_o = \frac{\lambda}{2S}\left[0.9102 + 6.8322\tanh\left(0.0109\frac{W}{h}\right)\right](pF \cdot \Omega) \quad (2)$$

$$L / Z_o = \frac{\lambda}{2S}\left[0.0103 + 0.0083\tanh\left(0.5665\frac{W}{h}\right)\right](nH / \Omega) \quad (3)$$

$$R / Z_o = 1.1040 - 1.2385\tanh\left(0.1036\frac{W}{h}\right) \quad (4)$$
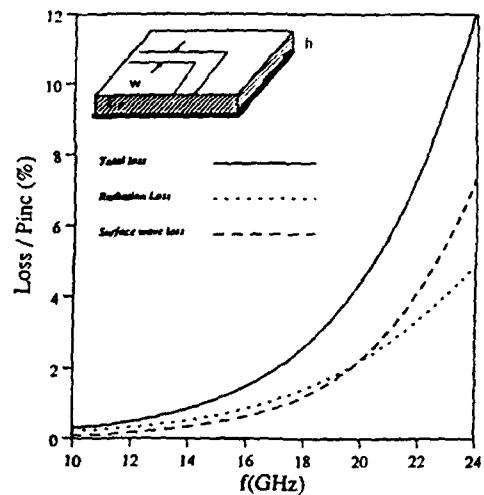
Fig. 4. Power losses versus frequency for right-angle bend discontinuity.
($\varepsilon_r = 10.2$, w = 24mil, h = 25mil)
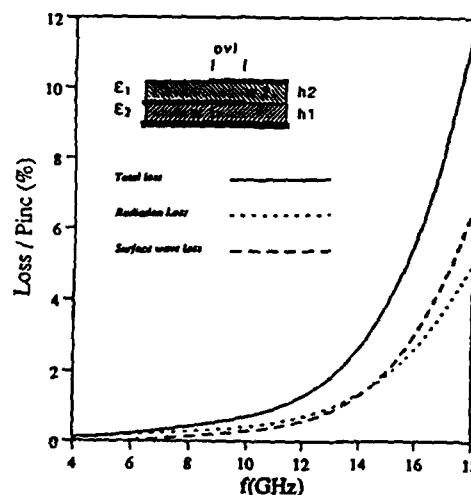
Fig. 5. Power losses versus frequency for overlay EMC lines.
($\varepsilon_1 = 2.2$, $\varepsilon_2 = 10.2$, h1 = 25mil, h2 = 25mil,
w1 = 42mil, w2 = 76mil , ovl = 83mil)

$\epsilon_r = 9.9, w = 24mil, h = 25mil, f = 24GHz$

y-directed current

x-directed current

incident wave

incident wave

Fig. 1. Integration plane for calculating radiation loss.

Computing time on IBM-3090 for matrix elements

GOAL:

To calculate the reaction integrals $Z_{ij}(m,n)$ between any two triangular basis functions

where

i,j : type of triangular basis function

m, n : relative distance index, m=[-50:50],n=[-50:50]

Difference part Z^d — 708 sec

Asymptotic part Z^a (Spatial domain) — 470 sec

Asymptotic part Z^a (Spectral domain) — 2832 sec

Total computing time $Z^a + Z^d$ — 1178 sec / 3540 sec

Fig. 2. Integration cylinder for calculating surface-wave loss.

microstrip discontinuity

d

microstrip discontinuity

Σ

$\varepsilon_r = 9.9$
h = 25.0 mils
w = 23.5 mils

$C_1 \cdot Z_0 = 0.64704$ (pF·Ω)
$C_2 \cdot Z_0 = 0.98044$ (pF·Ω)
$L / Z_0 = 0.01436$ (nH/Ω)
$R / Z_0 = 0.98381$

Fig. 3   Comparison of the Phase of the Open-End
Discontinuity between the Proposed Circuit
Model and the Measured Data

• Geometry of Microstrip Gap Discontinuity :

$\varepsilon_r = 9.9$
h = 25 mils
w = 25 mils
s = 10 mils

• Proposed Circuit Model :

$C_1 \cdot Z_0 = 0.83355$ (pF·Ω)
$C_2 \cdot Z_0 = 0.63088$ (pF·Ω)
$C_3 \cdot Z_0 = 0.50581$ (pF·Ω)
$L_1 / Z_0 = 0.01896$ (nH/Ω)
$L_2 / Z_0 = 0.02196$ (nH/Ω)
$R / Z_0 = 1.76299$
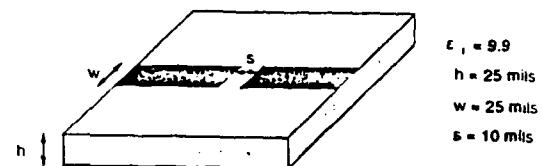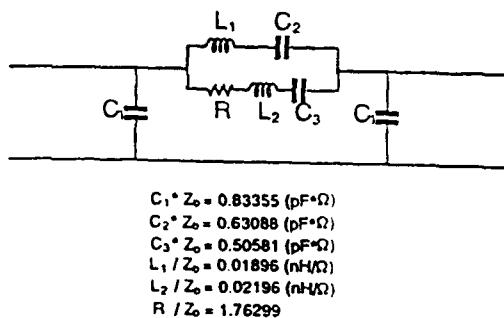
Fig. 4   Physical Dimension and the Equivalent Circuit
Model of a Microstrip Gap Discontinuity

Microstrip Open-End Geometry :
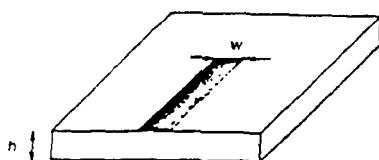
Proposed Circuit Model :
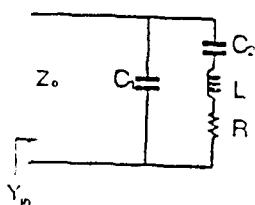
Fig. 1   Physical Dimension and the Equivalent Circuit
Model of a Microstrip Open-End Discontinuity.
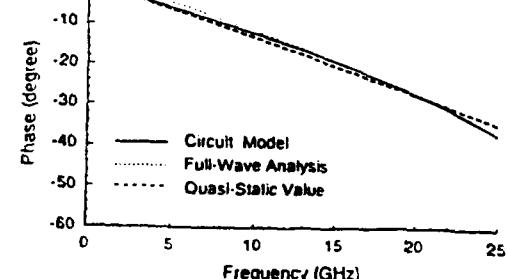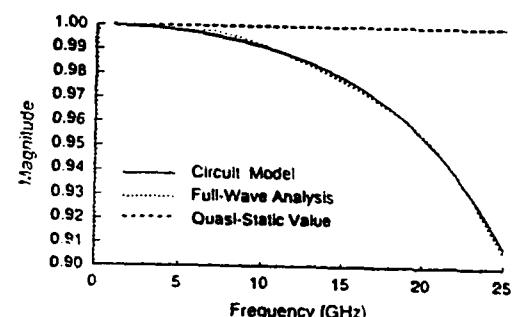
Fig. 2   Reflection Coefficient of a Microstrip Open-End
($\varepsilon_r = 9.9$, h = 25 mils , W = 25 mils)

Slot with Radial Open Ends

(c)

(d)

Fig. 5   S Parameters of a Microstrip Gap Discontinuity
($\epsilon_r = 9.9$, h = 25 mils , W = 25 mils , s = 10 mils)

(e)
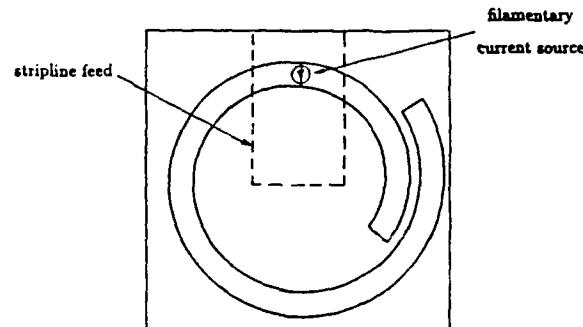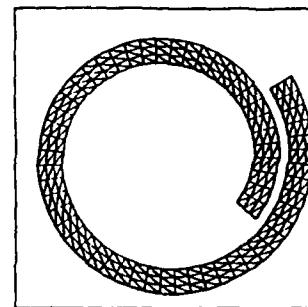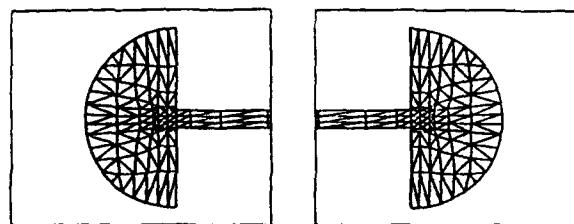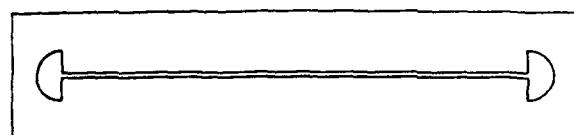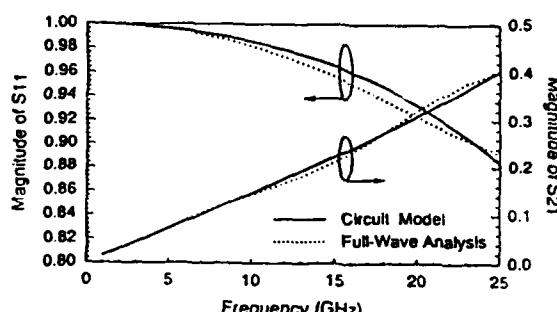
Potential Applications: Microwave and millimeterwave
aperture-type antennas whose geometries include
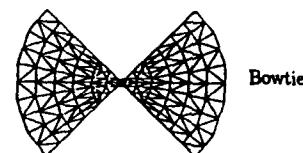
Bowtie
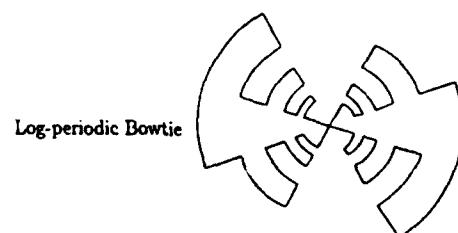
Log-periodic Bowtie

Archimedian Spiral

Distribution of Magnetic Current

$\frac{\lambda_1}{2\pi}$

$\epsilon_r = 1.0, \quad b/R = 0.2, \quad R/\lambda_1 = 0.847$
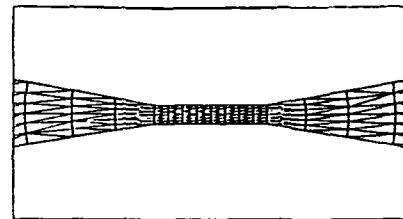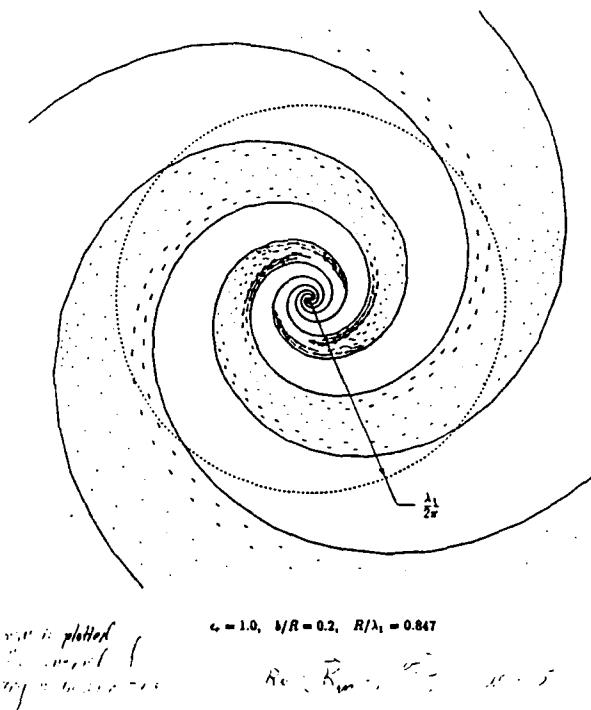
10° Bowtie Slot Antenna

(898 edges, 2000 mil radius, 40 mil gap)
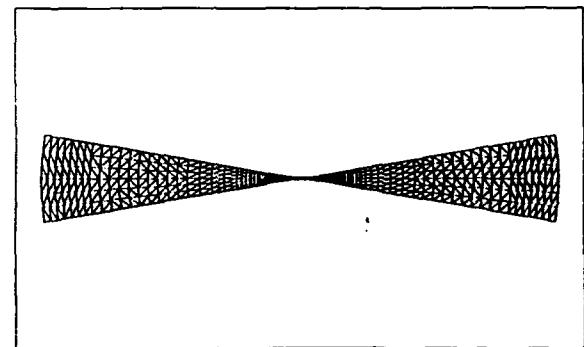
Equiangular Spiral Slot Antenna

$(\epsilon_r = 1.0, b = 300 \text{ mils}, R = 1000 \text{ mils})$

$(\rho = e^{a\phi + n\frac{\pi}{2}}, a = 0.3063/rad, n \in \{0,1,2,3\}, \quad 5 < \rho < 1000 \text{ mils})$

Equiangular Spiral Slot Antenna

$(\epsilon_r = 1.0, b = 300 \text{ mils}, R = 1000 \text{ mils})$

$(\rho = e^{a\phi + n\frac{\pi}{2}}, a = 0.3063/rad, n \in \{0,1,2,3\}, \quad 5 < \rho < 1000 \text{ mils})$

Normalized Power Pattern (dB) at 8.0 GHz

phi = 0 deg
phi = 90 deg

Axial Ratio Pattern (dB) at 8.0 GHz
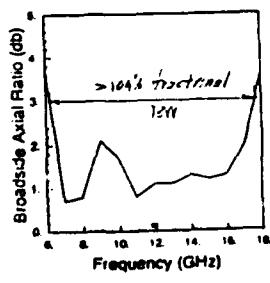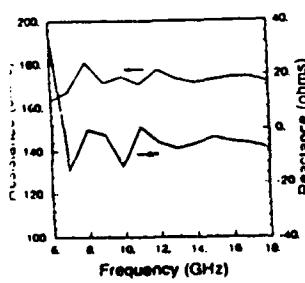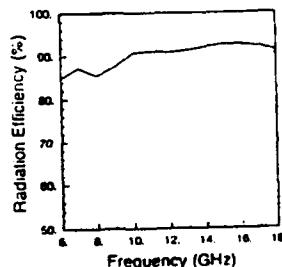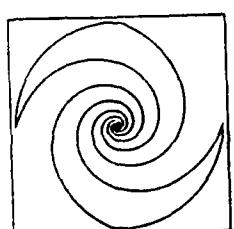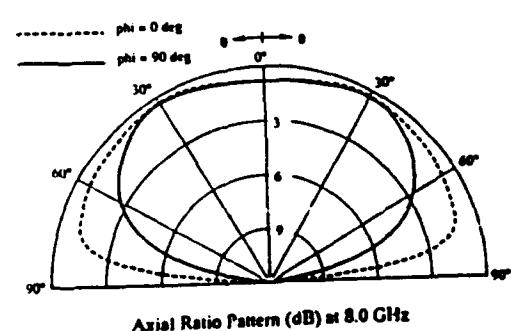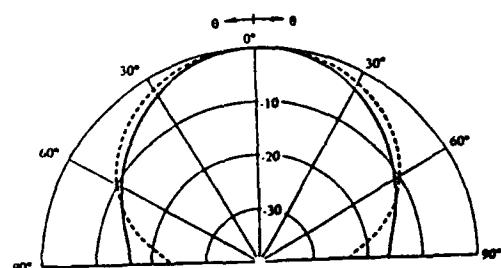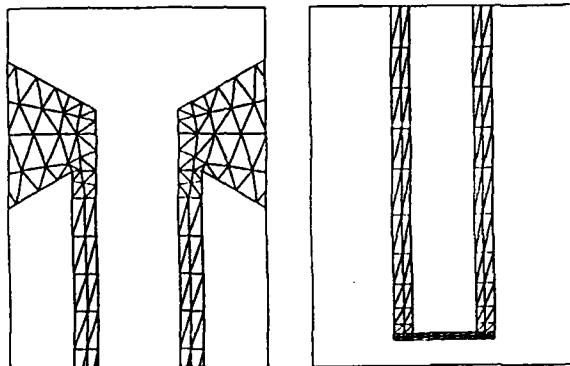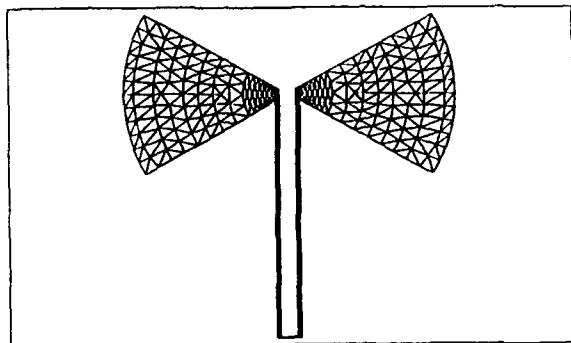
30° Bowtie Slot Antenna with 100Ω CPW Feed

(898 edges, 2000 mil radius)

Equiangular Spiral Slot Antenna

$$\rho = \exp\left[0.3063\left(\phi - n\frac{\pi}{2}\right)\right]$$
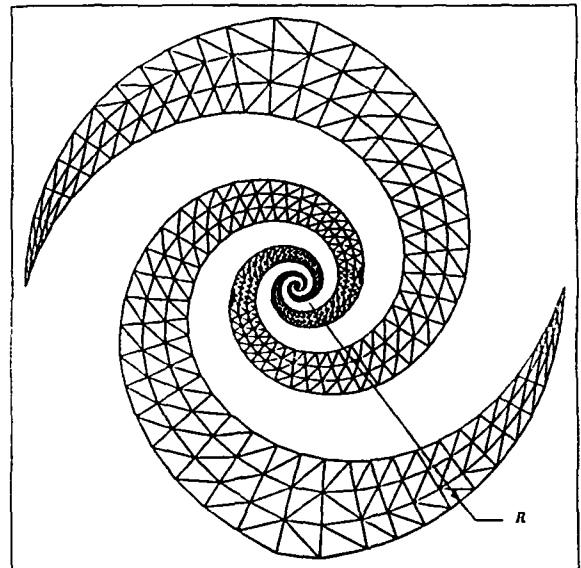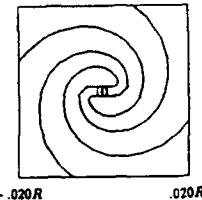
$$n = 0,1,2,3 \quad ; \quad 0.005R < \rho < R$$
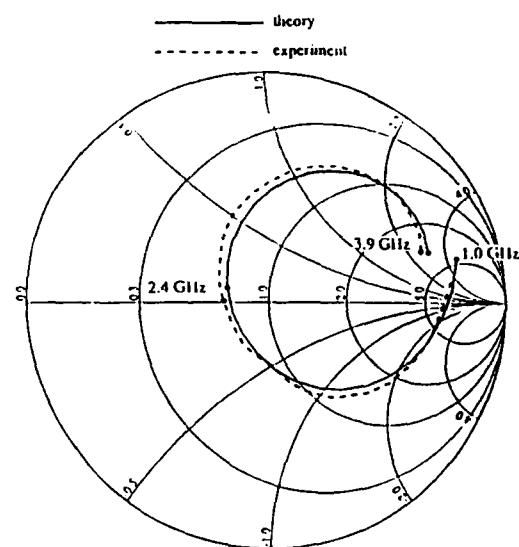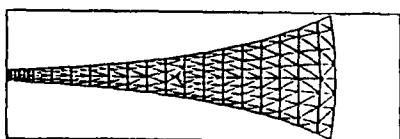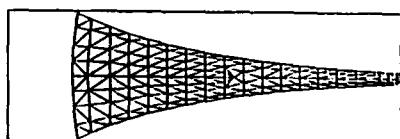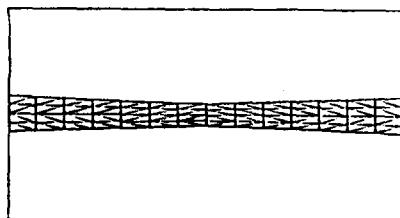

$-.020R$ \qquad $.020R$

R

Monofilar Archimedean Spiral Slot Antenna

$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$

$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$

Monofilar Archimedean Spiral Slot Antenna

$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$

$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$

Normalized Power Pattern (dB) at 12.6 GHz

------- phi = 0 deg

———— phi = 90 deg
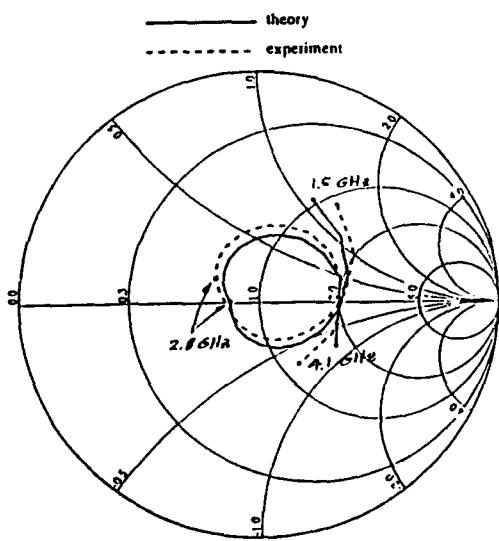

Axial Ratio Pattern (dB) at 12.6 GHz
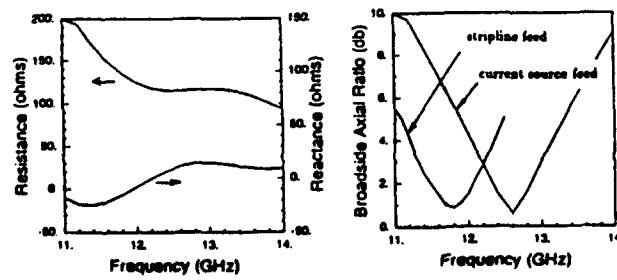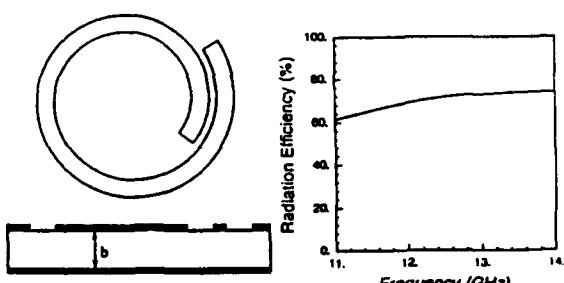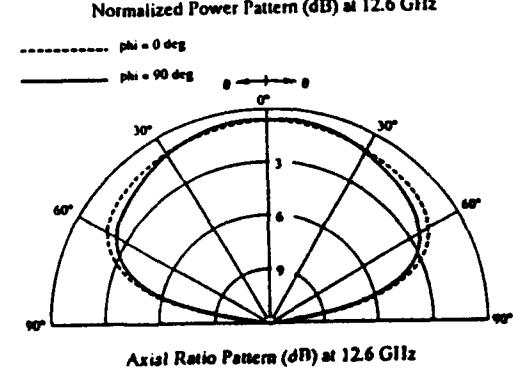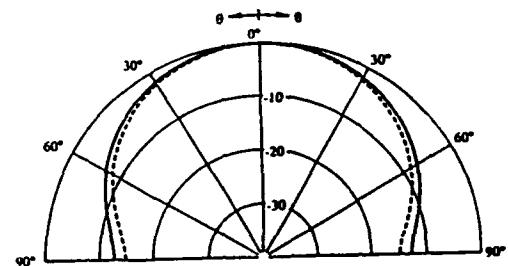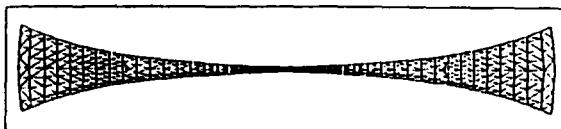
Exponentially-Tapered Slot Antenna

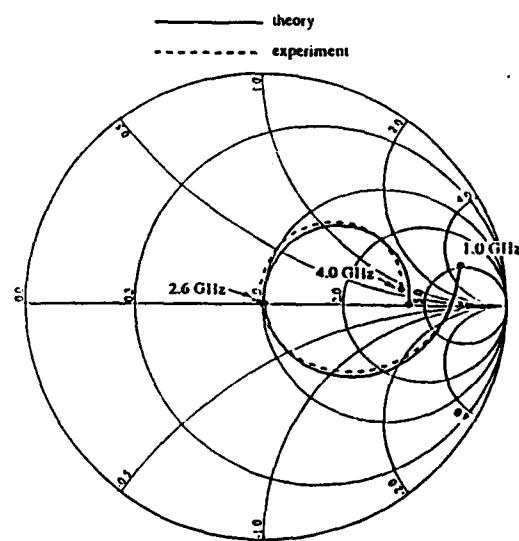(736 edges, 2000 mil radius, $w = 40 \exp(z/700)$ mils)

Reflection Coefficient for the Exponentially-Tapered Slot
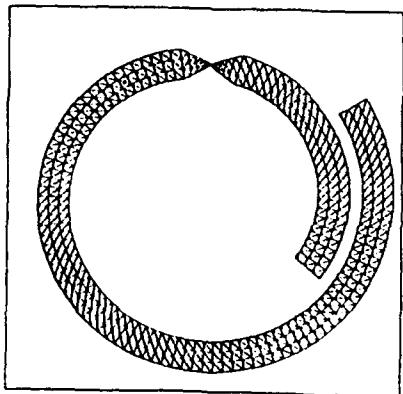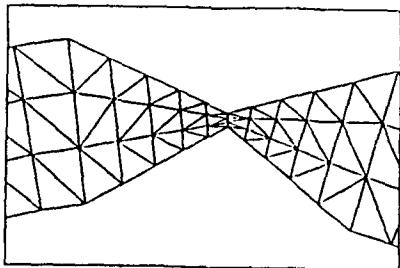
$Z_o = 50\Omega$

Reflection Coefficient, 30 Degree Bowtie w/ Feed

$Z_o = 100\Omega$

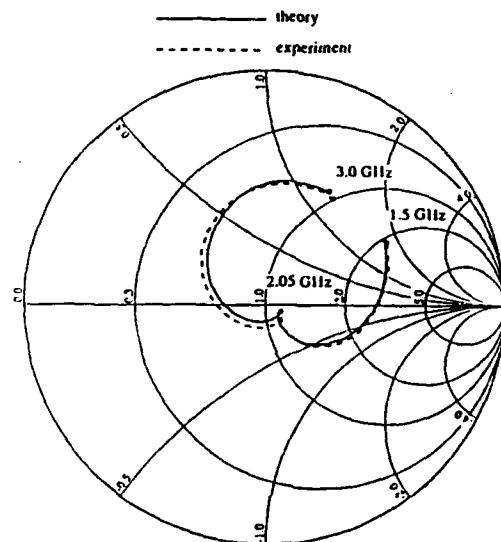Reflection Coefficient for the 10 Degree Bowtie

$Z_o = 50\Omega$

Monofilar Archimedean Spiral Slot Antenna

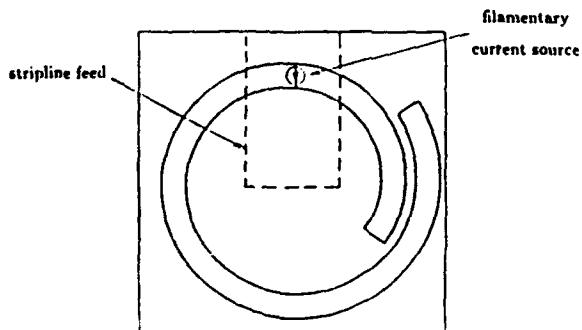$(\epsilon_r = 1.08, b = 1000 \text{ mils}, w = 254 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$

$(\rho = a\theta + \Delta, a = 57.5 \text{ mils/rad}, \Delta = 45 \text{ mils})$

Geometry of the Monofilar Archimedean Spiral

(989 edges, w=254 mils, r_{min} = 1004 mils, r_{max} = 1434 mils)

Reflection Coefficient for the Monofilar Spiral Slot Antenna

Z_o = 100 ohms

Monofilar Archimedean Spiral Slot Antenna

(Ref: Nakano, 1991 IEEE AP Symposium)

Monofilar Archimedean Spiral Slot Antenna

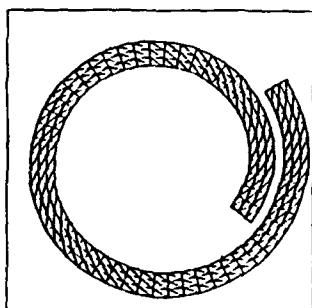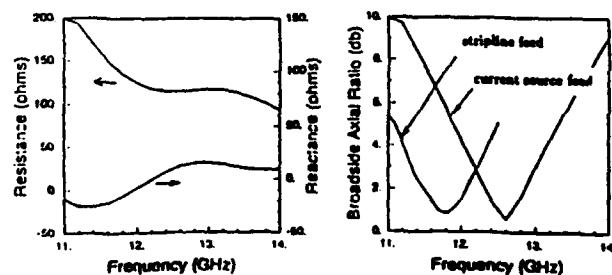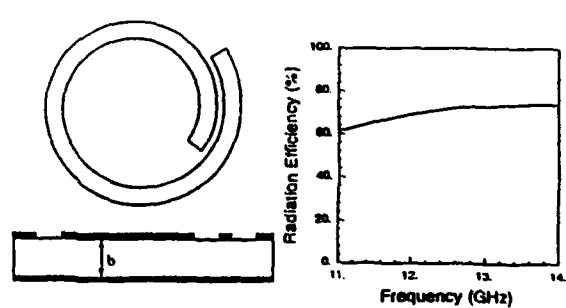$(\epsilon_r = 2.2, b = 500 \text{ mils}, w = 40 \text{ mils}, \theta_{st} = 5.81\pi, \theta_{end} = 8.188\pi)$

$(\rho = a\theta + \Delta, a = 9.055 \text{ mils/rad}, \Delta = 7.087 \text{ mils})$

Standard integral equation approach

• For a given β, ω is found by solving a <u>non-linear</u> eigenvalue problem

• The solution of this problem requires the search of the zeros of a ω-dependent matrix, which must be recalculated at each step in the search

• The search of the zeros is a time consuming procedure especially when:

- the propagating frequencies of many modes are required

- no estimate of the expected values is available

- quasi-degenerate modes must be resolved

• The troubles of the integral equation approach derive from the fact that $\underline{G}$ is a transcendental function of ω

• In each elementary waveguide the e.m. field can be considered as generated by equivalent current and charge sheets placed at the magnetic walls

$$J_s(x,y) = \pm u_x \times H(x,y) \qquad \rho_s(x,y) = - \nabla \cdot J_s(x,y) / j\omega$$

where $\nabla = u_x \dfrac{\partial}{\partial x} + u_y \dfrac{\partial}{\partial y} - j\beta\, u_z$

• The electric field can be expressed as:

$$E(x,y) = - \nabla \int F(x,y;x',y';\beta)\, \rho_s(x',y')\, dy' - j\omega \int_{\text{magnetic walls}} \underline{G}(x,y;x',y';\omega,\beta) \cdot J_s(x',y')\, dy'$$

where the integrals represent the scalar and the vector potentials in the Coulomb gauge, and F and $\underline{G}$ are the pertinent Green's functions. They are the solution of the problem:

$$\nabla \cdot \varepsilon \nabla F = - \delta(x-x')\, \delta(y-y')$$

$$\nabla \times \mu^{-1} \nabla \times \underline{G} - \omega^2 \varepsilon \underline{G} = \underline{1}\, \delta(x-x')\, \delta(y-y') - \varepsilon \nabla \nabla' F$$

with appropriate boundary conditions.

• Functions F and $\underline{G}$ can be determined analytically in each region.

• Enforcing the continuity of $u_x \times E$ at the interfaces between the elementary waveguides a system of integral equations is obtained.

• The problem is put into algebraic form using the method of the moments, with base functions suitable to represent both the components of the unknown currents (J_s) and the related charges (ρ_s) and taking into account the singularities at the metallic and the dielectric edges.

Departement of Electronics
University of Pavia
via Abbiategrasso 209, 27100 Pavia (Italy)

Microwave Group

G. Conciauro   (responsible)
P. Arcioni
M. Bressan
P. Gamba
L. Perregrini

Present activities on numerical modelling of passive structures :

• Quasi planar waveguides and transmission lines
• Quasi planar 3 D circuits
• Cavities resonators

Development of non conventional approaches aiming at

• flexibility
• accuracy
• efficiency

The following presentation aims at giving an idea of the approach under investigation.

A new method for numerical modelling of general quasi-planar waveguides

M. Bressan, G. Conciauro, P. Gamba

(to be presented at U.R.S.I. Symposium on Electromagnetic Theory, Sydney, 17-20 August 1992)

• Most of the quasi-planar millimeter wave and optical waveguides can be segmented into side-coupled uniformly stratified rectangular waveguides.

---- magnetic wall
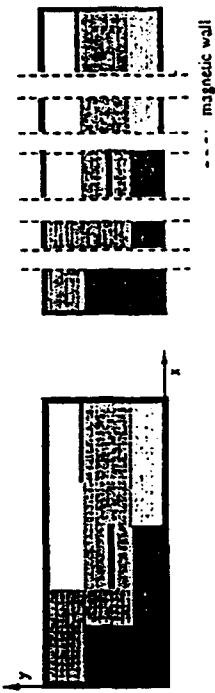
• For any propagating mode, the e.m. field in the structure can be expressed as:

$$E = E(x,y)\, e^{j(\omega t - \beta z)} \qquad H = H(x,y)\, e^{j(\omega t - \beta z)}$$

• The algorithm permits one to calculate the dispersion $(\omega - \beta)$ diagram and the modal fields.
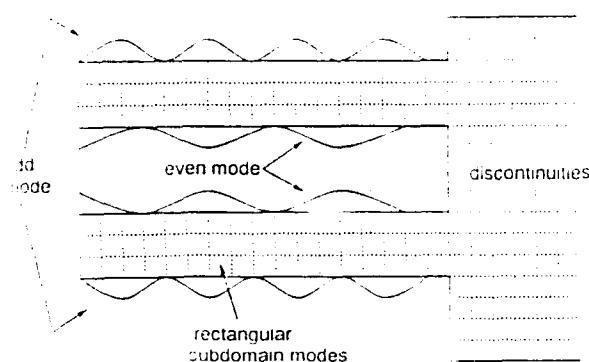
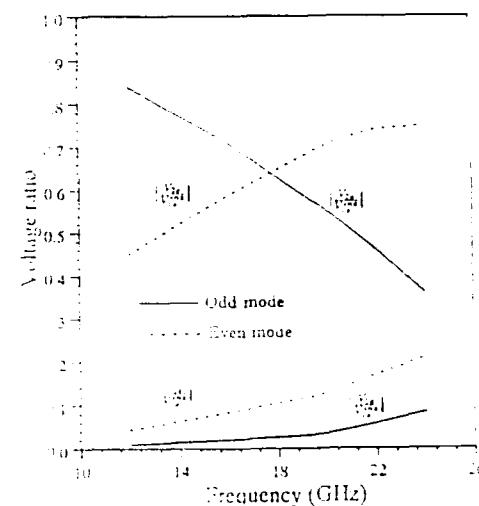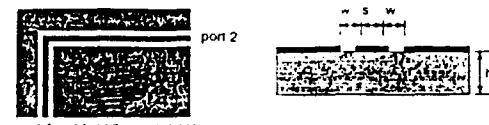# PARTECIPANTS' CONTRIBUTIONS

Fig. 2. Layout of expansion modes for magnetic currents on slots

Fig. 3. The slot voltages of even modes and odd modes in a CPW right-angle bend (ϵ_r=9.9, w=24mil, s=24mil, h=25mil)

Computational Performance

of the Arbitrary-Geometry Modeling Software

on the IBM 3090

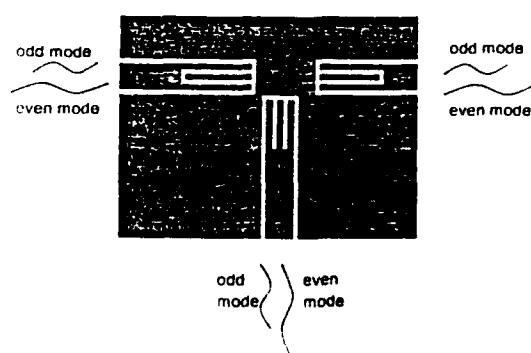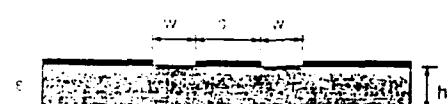| antenna elements	no. of basis functions	execution time per frequency (seconds)	storage required (megabytes)
Slot Dipole	200 - 500	50 - 300	.05 - 4
Bowtie Slot	400 - 800	250 - 500	2.5 - 11
Monofilar Spiral	500 - 800	300 - 500	4 - 11
Archimedian Spiral	700 - 2000	450 - 2000	8 - 64
Equiangular Spiral	1000 - 3000	800 - 3500	16 - 150

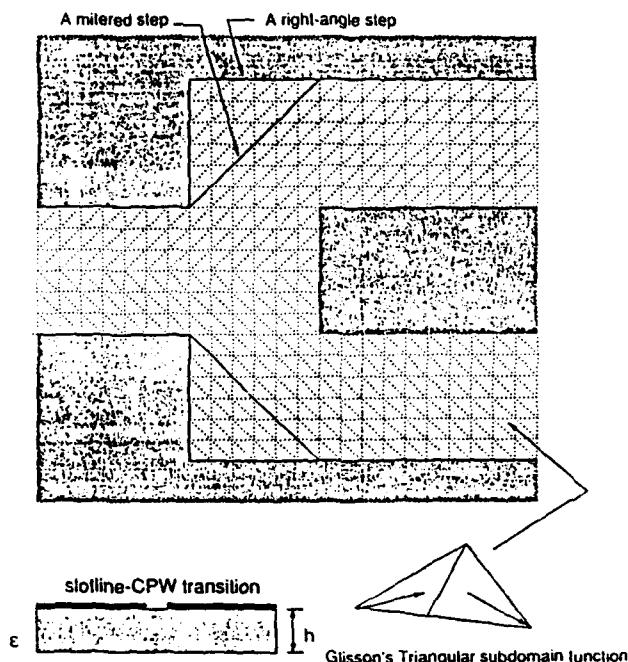Fig. 1. Asymmetric coplanar waveguide discontinuities

Fig. 1. Slot-to-CPW transition
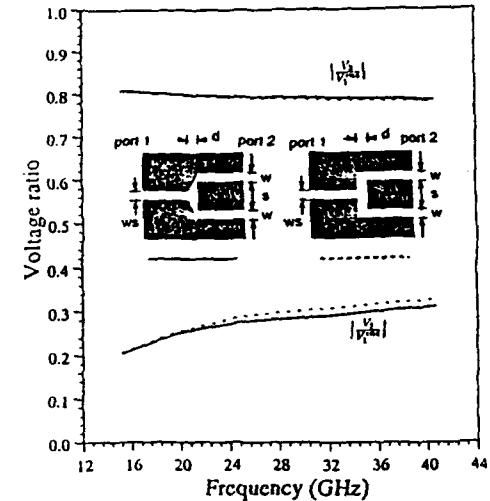
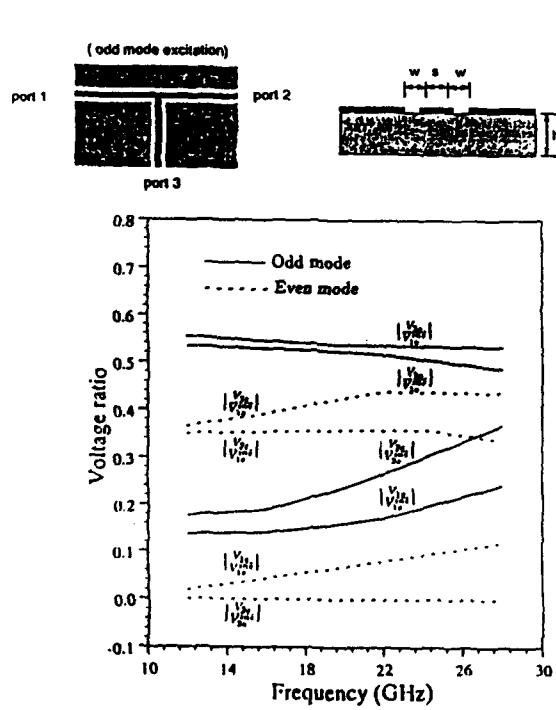Fig. 2. The slot voltages in a slotline-CPW junction (ϵ_r=9.9, ws=0.24mm, w=0.24mm, s=0.24mm, h=0.25mm, d=0.24mm)

Fig. 4. The slot voltages of even modes and odd modes in a CPW tee junction (ϵ_r=9.9, w=24mil, s=24mil, h=25mil)
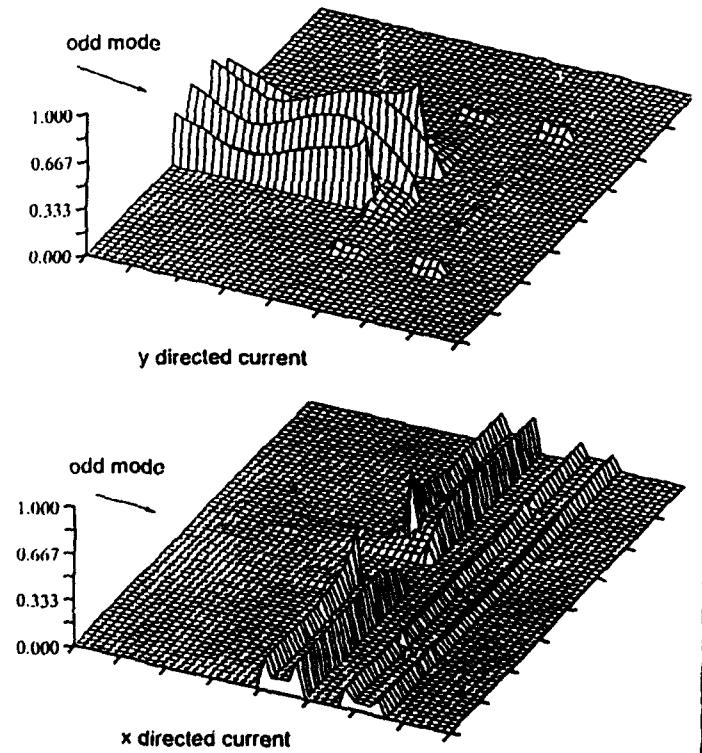
Fig. 5. Magnitude of equivalent magnetic current in a tee junction

Microstrip loaded Inset Guide

a T. Rozzi et al.,
   IEEE MTT, 1992

a = 10.16 mm
b = 10.42 mm
w = 2   mm

A MILLIMETER WAVE SIX-PORT REFLECTOMETER

S.ABOUCHAHINE, B.HUYART, E.BERGEAULT ,
L.JALLET

TELECOM PARIS , 46,RUE BARRAULT, 75634 PARIS
FRANCE
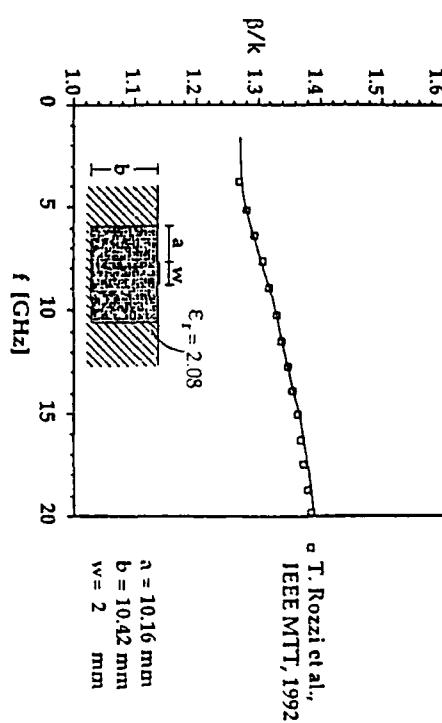
Abstract

This paper presents a six- port reflectometer in the W band
using Schottky diodes as power detectors. A linearization procedure and
AC detection technique are used to improve the measurement accuracy.
Morever, a calibration procedure requiring two standards is described.

Hybrid integral equation/mode matching approach

- $\underline{\underline{G}}$ can be represented as:

$$\underline{\underline{G}}(x,y,x',y',\omega,\beta) = \underline{\underline{G}}(x,y,x',y',0,\beta) - \omega^2 \sum_m \frac{\underline{g}_m(x,y,\beta)\,\underline{g}_m(x',y',\beta)}{[\omega^2 - \omega_m^2(\beta)]\,\omega_m^2(\beta)}$$
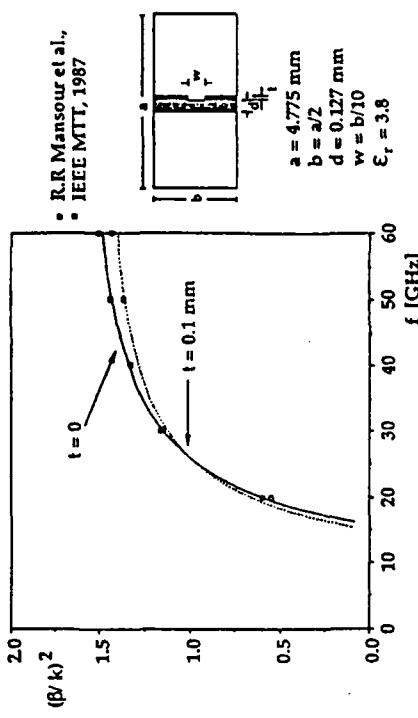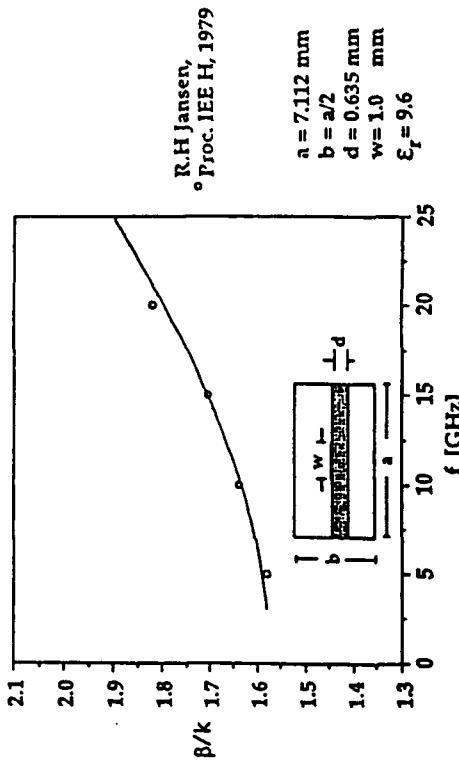
where ω_m and g_m represent the electric fields and the cutoff frequencies of the modes of the elementary waveguides. They are the eigenvalues and the eigenvectors of the problem:

$$\nabla \times \mu^{-1} \nabla \times g - \omega^2 \epsilon\, g = 0$$

with suitable boundary conditions and normalization.

- Due to its fast convergence (like ω_m^{-4}) the series can be truncated without relative convergence problems.

- After truncation $\underline{\underline{G}}$ is represented by a rational function of ω.

- Introducing a set of auxiliary variables related to the mode amplitudes in the elementary waveguides, the problem arising from the method of the moments can be converted into a *linear eigenvalue problem*, involving *frequency independent matrices*.

- Since the variables involved represent both the fields at the interfaces and the amplitudes of some modes, the described technique can be viewed as an hybrid between the boundary integral equation method and the mode matching technique.

Unilateral Finline

- R.R Mansour et al.,
- IEEE MTT, 1987

a = 4.775 mm
b = a/2
d = 0.127 mm
w = b/10
ϵ_r = 3.8

Suspended Strip Line

R.H Jansen,
° Proc. IEE H, 1979

a = 7.112 mm
b = a/2
d = 0.635 mm
w = 1.0 mm
ϵ_r = 9.6

5

FREQUENCY DEPENDENT ANALYSIS OF PACKAGING PROBLEMS
USING THE THREE-DIMENSIONAL INTEGRAL EQUATIONS TECHNIQUE

H. GHALI*, M. DRISSI*, J. CITERNE* and V. FOUAD HANNA**

*Laboratoire Composants et Systèmes pour Télécommunications
URA 834, INSA, 35043 RENNES, FRANCE
**CNET/PAB, 92131 ISSY LES MOULINEAUX, FRANCE

ABSTRACT

The main problems associated with packaging technique for microwave and millimeter integrated or monolithic circuits are the shielding effects, discontinuity and chip connections characterization. In this paper, a dynamic analysis, based on the three-dimensional integral equations technique associated with the method of moments is presented for the characterization of shielded discontinuity problems. It uses the three-dimensional Green's function for a general current distribution inside a shielded box. The theory of loaded scatterers is also used to take into account the effect of existing localized and distributed active devices or passive loads. Using this last theory, a virtual matched load is simulated, numerically, by introducing localized absorbing loads at circuit sections corresponding to output ports. Consequently, a scattering parameters extraction technique, based on the use of these matched terminations, is developed, and hence the [S] parameters of the studied discontinuity are determined from the knowledge of only current or electric field maxima which are often accurately determined under good matching conditions. The shielding effects on a wide range of planar and three-dimensional discontinuities are given. The behaviour of a typical chip connection as a strip air bridge is also studied and the frequency dependent characteristics of these discontinuities are presented.

Single Side Source Grounded Configuration

Guido Baumann
SEL ALCATEL
Mannheimerstr. 22b, 7530 Pforzheim, Germany

ABSTRACT

This paper presents an oscillator design using a new concept of a FET model, which allows to characterize the single side source grounded configuration of a FET, and therefore to predict the oscillator frequency in the mm-wave range very precisely.
The freerunning oscillator was fabricated on a GaAs chip. At 36 GHz an output power of 14 dBm, and an external quality factor of greater than 100 was achieved.

SUMMARY

For an accurate oscillator design in the mm-wave range the modeling of the active device is essential for a successful design. Especially oscillators are often layouted in a single side grounded source (SSG) configuration, that means only one source pad is grounded, while the second one is not connected.

Depending on the gatewidth and the dimensions of the FET, distributed effects for low noise MESFETs have to be regarded above approximately 50 GHz /1/, /2/. This assumption is only valid if both source pads are grounded.

If the FET is used in a single side grounded source configuration, the FET should be regarded as distributed even above 15 GHz. To our knowledge the effect of SSG has not been regarded yet by any commercial CAD tool.

Especially for oscillator design the FET is often used in a SSG configuration /3/ to reach the required unstability, otherwise both source connections will have symmetrical networks which will increase the area on a chip. The difference in the measured S-parameters between single side source grounded (SSG) and double side source grounded (DSG) configuration is shown in Fig. 1. Above 15 GHz the S-parameters S_{12} and S_{11} show great differences.

From the layout orientation of the 4-finger MESFET TEG4-50-25, Fig. 2, a modular distributed model was deduced, Fig. 3. A 4-finger MESFET with a 0.25 μm gatelength and a total gatewidth of 200 μm from TEG' is used. The geometrical dimensions of the chip and source configuration are directly implemented as microstrip elements. The source pad which is not grounded in the SSG configuration is described by a tapered open line. The distributed model provides 3-port S-parameters for SSG configuration. Fig. 4 indicates the excellent agreement of the modeled and measured S-parameters for SSG configuration.

The FET model is also able to describe the reverse channel operation. S-parameter measurements of the MESFET were performed using an inhouse developed testfixture with TRL calibration standards for microstrip measurements up to 40 GHz.

The basic concept of the Oscillator is a reflection type oscillator with the output at the drain side /4/, /5/. In Fig 5, the layout of the 28 GHz oscillator is shown. The FET is positioned at the edge of the chip and only one sourcepad is extended with a microstrip transmission line to provide the required unstability.

The GaAs oscillator has a chip size of 3x3 mm and is fabricated on a 150 μm GaAs substrat. The performance of the 36 GHz freerunning Oscillator with U_{ds}=5V and I_{ds}=30mA has a power output P_{out} = 14 dBm, an external quality factor Q_{ext} = 100.

shown in Fig. 6 and Fig. 7, respectively.
Using a common FET model for this oscillator design, the simulated and measured resonance frequency had an offset of about ±3 GHz. With the new distributed FET model the predicted resonance frequency of the freerunning oscillator was within a range of ± 200 MHz.

CONCLUSION

This paper shows that the FET has to be regarded as distributed above 15 GHz if the FET is used in a SSG configuration. The deduced model describes very accurately the SSG configuration and uses only lumped elements and microstrip lines which makes it practicable for online simulation. This leads to a higher accuracy in oscillator design. The predicted resonance frequency of the freerunning oscillator lies within a range of ±200MHz from the design frequency. Oscillators designed with common FET models had their resonance frequency within a range of ±3 GHz. For nonlinear modeling the intrinsic FET will be changed to a large signal model to provide also the possibility of nonlinear circuit design in SSG-configurations.
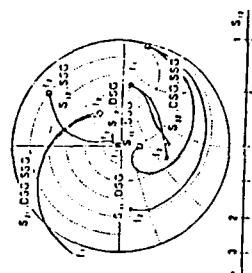

Fig. 1: Measured S-parameter in double side grounded (DSG) and single side grounded (SSG) configuration of a MESFET TEG4-50-25: $f_1 = 1$ GHz, $f_2 = 40$ GHz.
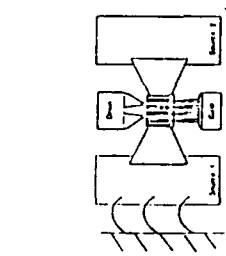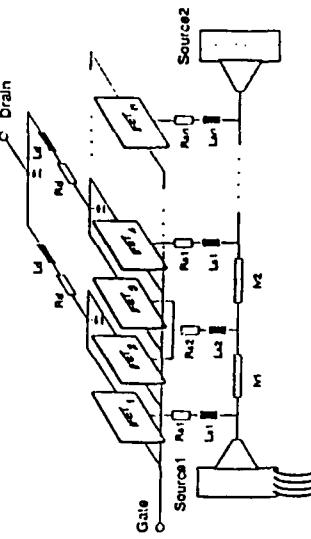

Fig. 2. Layout of the 4-finger MESFET in single side grounded configuration (SSG).

Fig. 3: Modular FET model with distributed source configuration for T-Gate FET in SSG configuration.
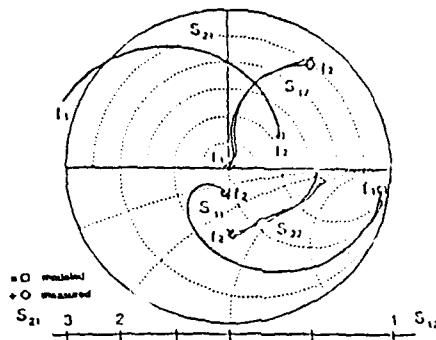
1

2

Fig. 4: Measured and modeled S-Parameter in single side grounded configuration (SSG).
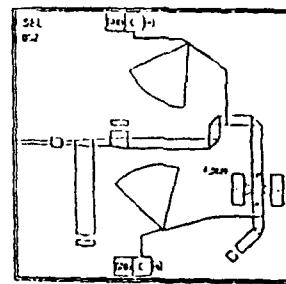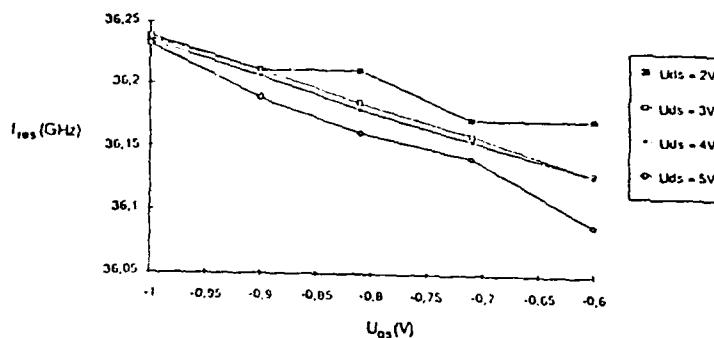

Fig. 5: Layout of the Oscillator.

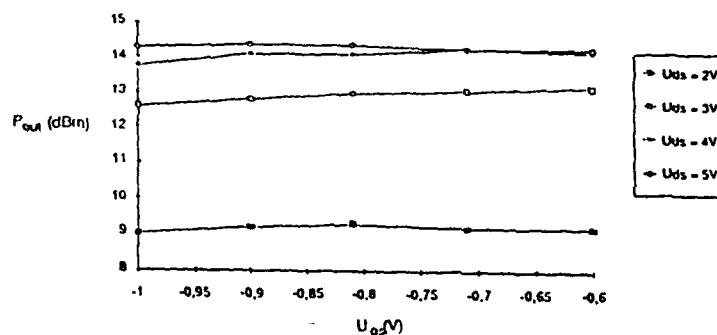Fig. 6: Resonance frequency vs gate- and drain voltage.

Fig. 7: Output power vs gate- and drain voltage.

References:

/1/ W. Heinrich, "Limits of FET Modelling by Lumped Elements", *Electronics Letters*, vol. 22, no. 12, S. 630-632, Juni 1986

/2/ H.P. Feldle, "Klein- und Großsignal-CAD-Modelle von GaAs MESFETs mit verteilten Elementen", Dissertation Universität Karlsruhe, 1987

/3/ G. Baumann, R. Hierl, "Modular FET model with distributed source configuration for a single and double side source grounded MESFET", *Electronics Letters*, vol. 27, no. 13, S. 1128-1129, Juni 1991

/4/ K. Kurokawa, "Some Basic Characteristics of Broadband Negative Resistance Oscillator Circuits", *The Bell System Technical Journal*, Juli- August, 1969

/5/ George D. Vendelin, Design of Amplifiers and Oscillators by the S-Parameter Method, John Wiley and Sons, Inc., 1982